FEBRUARY 1958 VOLUME 2 (1958) No. 2

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NUCLEAR INSTRUMENTS

A JOURNAL ON ACCELERATORS, INSTRUMENTATION AND TECHNIQUES IN NUCLEAR PHYSICS

EDITOR K. SIEGBAHN - UPPSALA

CERN MEETING

on

NUCLEONIC INSTRUMENTATION

for

HIGH ENERGY PHYSICS

edited by

G. VON DARDEL

NORTH-HOLLAND PUBLISHING COMPANY - AMSTERDAM

NUCLEAR INSTRUMENTS

A journal on accelerators, instrumentation and techniques in nuclear physics

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Nuclear Instruments is published in bi-monthly issues. The subscription price of a volume of 360 pages is $ 17.00 (130 s), postage free.

Subscriptions should be sent to the publishers, North-Holland Publishing Company, P.O.Box 103, Amsterdam or to any subscription-agent. Subscriptions for U.S.A. are to be sent exclusively to Interscience Publishers, Inc., 250 Fifth Avenue, New York, 1 N.Y.

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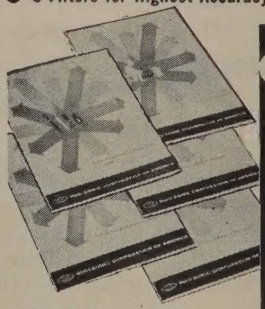

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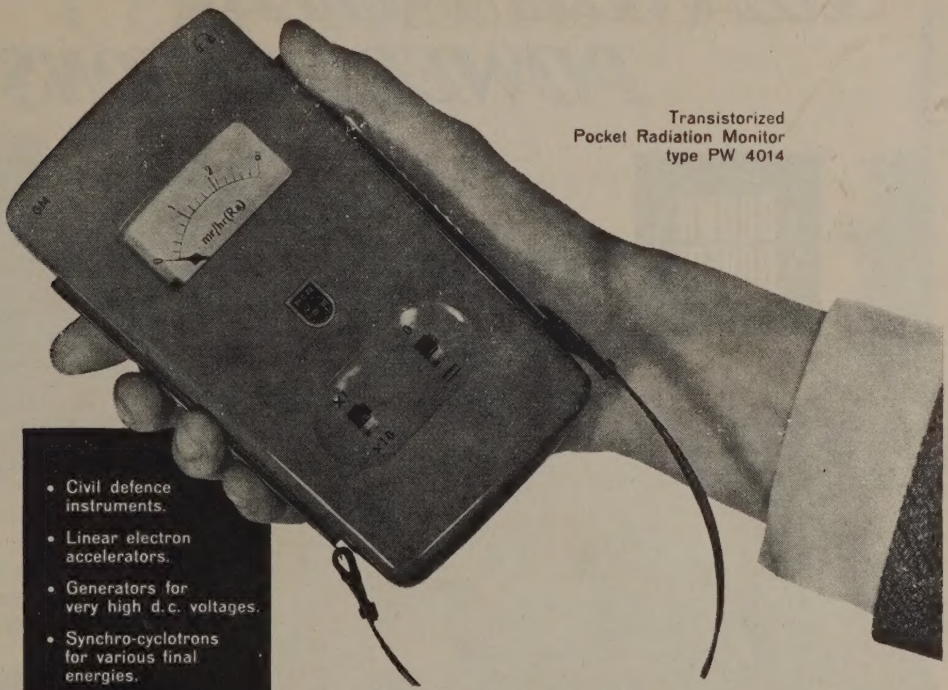

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OPENING SPEECH
FOR MEETING ON NUCLEONIC INSTRUMENTATION

C. J. BAKKER

Director General, CERN, *Geneva*

During the next few years a number of high energy accelerators will come into operation. The 600 MeV CERN synchrocyclotron is now running and the 25 GeV protonsynchrotron is scheduled for operation in 1960 or 1961. 1 GeV electronsynchrotrons are built in Rome and Lund. The 3 GeV machine in Saclay should operate in a few months. A 7 GeV machine is under construction in Harwell. Other high energy machines are being planned.

These costly machines will have to be provided with adequate instrumentation needed for the experiments. Much of this instrumentation is of a nature where a close contact and cooperation between the instrumentation groups in the various institutes would be very useful. Unnecessary duplication of efforts could thus be avoided. Early contacts between the instrumentation groups will also facilitate the future adaptation problem when research teams will perform experiments with the CERN machines using both its own equipment and instruments lent by CERN.

Informal meetings of this kind between Harwell, Saclay, Liverpool and CERN have taken place periodically during the last two years. The present meeting is an extension of these previous meetings to englobe other institutes as well. It should also serve to establish preliminary contacts between the instrumentation groups of the different European nuclear research centres. The participants to the meeting have been proposed by the delegations of the different countries to CERN. The number of participants to the meeting has been severly limited so as to allow intimate and informal discussions which are essential in this highly technical field. It was then unavoidable that some of the institutes which would be interested to participate could not be invited. This disadvantage will not be serious if discussions of general interest are convered by adequate minutes with a fairly wide distribution. It is therefore envisaged to publish the proceedings of the meeting either as a CERN report or in a scientific journal with a more widespread circulation.

I hope that the three days of the meeting will prove of great interest to the participants and I trust that the contacts which are established will be of value for the future.

EUROPEAN ORGANIZATION FOR NUCLEAR RESEARCH

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CERN Meeting on Nucleonic Instrumentation for High Energy Physics

Meyrin – Geneva

30 September – 2 October 1957

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SESSION I.1

SOME LIMITATIONS AND PRINCIPLES OF NANOSECOND TIME MEASUREMENT IN NUCLEAR PHYSICS

H. DE WAARD

Natuurkundig Laboratorium, Rijksuniversiteit, Groningen

1. Introduction

Already before the second world war extensive use was made of *microsecond* pulse techniques in circuits associated with gas filled nuclear radiation detectors. Of course, the fast development of these techniques for radar and television during and after this war proved of great importance also to nuclear physics. When much faster types of detectors, such as scintillation and Čerenkov counters,—in which gas amplification is replaced by secondary emission amplification in a vacuum—came into use less than a decade ago, the existing techniques were quickly incorporated in the circuits associated with the new detectors, but they were soon found to be incapable of exploiting the very short time resolution that could be obtained with such counters. From the shops of the nuclear physicists themselves new techniques, which we now call *nanosecond* pulse techniques, then started to emerge. These were especially adapted to the new methods of detecting nuclear events. At the moment one may observe that such techniques are gaining importance even in other fields. Indeed, it seems that nuclear physics may give new impetus to the very fields from which it first derived its basic requirements in pulse technique.

In many cases, the distinction *nano-* vs. *micro*second techniques, implying a shortening of manageable time intervals by a factor of a thousand may seem a trifle ambitious: in quite a few circuits the improvement hardly exceeds a factor of ten—take for instance time resolution in fast scalers. Yet, ultimate time resolutions obtainable with scintillation counters may be better by a factor of one thousand as compared with counters that use gas amplification and therefore the term nanosecond techniques appears to be justified, if only to spur physicists on to improve further the nanominute or picohour resolutions obtained presently.

Why do we strive so after ever shorter time resolution? The answer is easy to find by looking at problems in nuclear physics that are fashionable right now. As an example, take lifetimes τ of very short lived nuclear levels. These may be determined by measuring the—statistically distributed—time intervals between the emission of particles exciting and de-exciting the levels, employing the well-known delayed coincidence technique. In the region $10^{-10}\,\mathrm{s} > \tau > 10^{-12}\,\mathrm{s}$ that can hardly be reached with presently available instruments, theory makes interesting predictions that we should like very much to verify. Another example is provided by the time of flight experiments with fast particles that are so rapidly gaining popularity and that are undoubtedly of special importance to many of you here. In this case better time resolution means better energy resolution. When such time of flight experiments are performed with a pulsed source—for instance a cyclotron, where phase bunching[1,2]) automatically gives pulsed operation, or a Van de Graaff or cascade generator in which the beam is swept across the

[1]) R. Grismore and W. C. Parkinson, Rev. Sci. Instr. 28 (1957) 245.

[2]) K. G. Malmfors, J. Kjellman and A. Nilsson, Nuclear Instruments 1 (1957) 186.

target[3])—delay fluctuations in the detector, and the finite width and possible jitter of the pulses from the source may decrease the obtainable resolution. As long as the particles from the source are strictly mono-energetic, the influence of these effects may in principle be reduced by increasing the length of the flight path, but in practice available intensities often limit flight paths to such lengths that the timing uncertainties form the chief limiting factors.

Of the factors that limit time resolution, some are of a fundamental physical nature—for instance the finite fluorescence decay time of luminescent materials in scintillation counters—others appear to be of a more technical nature—such as transit time spread in photomultipliers. In general, it may further be said that in quite a few cases phenomena in the radiation detector itself nowadays set limits of performance rather than the electronic circuits associated with it. This is clearly evidenced, for instance, by the improvement in performance of fast coincidence circuits when driven by artificial pulses instead of counter pulses. (By the way, this ought to show how misleading it can be to judge instrument performance with artificial pulses—though of course a source of such pulses can be extremely handy as a means of checking certain circuit characteristics quickly.)

In view of its primary importance we will therefore first look into factors affecting time resolution of fast radiation detectors. After that we will discuss some operations that may be carried out by fast circuits on the pulses these detectors produce, restricting ourselves here to methods for measuring short time intervals.

2. Time Resolution in Scintillation and Čerenkov Counters

In these counters a—mostly transparent—material, penetrated by rays or particles, becomes luminescent. As many photons as possible are collected from the phosphor on to the photocathode of a photomultiplier tube, where some of them liberate photo-electrons. These enter the secondary emitting dynode structure of the tube, starting avalanches of secondary electrons that result in sizeable current pulses at the anode.

The output pulses exhibit fluctuations both in *size* and *rise-time*. Suppose they are fed into a circuit that reacts as soon as they exceed a certain threshold value. Then, both size and rise-time spread will contribute to an apparent delay-spread in the circuit. Often the influence of fluctuations in size can be minimized by incorporating a pulse height selector that activates the instrument only when the pulses fall within a narrow predetermined pulse height interval. Therefore we will restrict ourselves here to rise-time fluctuations. Factors that influence rise-time are:

(1) fluctuations in delay of photon emission by the phosphor.

(2) fluctuations in the photon collection time, resulting from the finite size of the phosphor and possibly from a light guide between phosphor and photocathode.

(3) fluctuations in transit time of: (a) photo-electrons, (b) secondary electrons in the multiplier.

We will estimate first how factors (1) and (3) contribute to delay spread, neglecting (2) for the moment†.

In Čerenkov counters fluctuations of type (1) will only result from the finite time of flight of super-photic particles in the phosphor, the emission of Čerenkov light being instantaneous on passage of such particles. In general, such fluctuations will be negligible compared with those of type (2) and (3).

In scintillation counters, photon emission from the phosphor decays exponentially with time, and the decay rate may for our purposes be described by a single decay time τ. The average number of photo-electrons produced at the

[3]) W. Weber, C. W. Johnstone and L. Cranberg, Rev. Sci. Instr. **27** (1956) 166.

† Arguments similar to what follows have already been presented by Morton, Nucleonics, **10** (1952) 39. A more extensive treatment of scintillation counter statistics may be found in ref.[5]) and very recently in: S. Colombo, E. Gatti and M. Pignanelli, Nuovo Cimento, **5** (1957) 1739.

photomultiplier cathode within t seconds after excitation of the phosphor is then

$$n = m\,(1 - \exp(-t/\tau)) \approx m\,t/\tau \quad \text{if} \quad t \ll \tau \qquad (1)$$

m being the average total number of photo-electrons per scintillation.

The r.m.s.-deviation of the time of emission of the n–th photo-electron can be approximated by

$$\sigma_p \approx \tau\sqrt{n}/m \quad \text{if } t \ll \tau \quad (n \ll m)\ . \qquad (2)$$

Post and Schiff[4]) have given a more exact expression for σ_p, which is also valid if no longer $n \ll m$. It turns out that σ_p exceeds (2) increasingly as n/m approaches 1.

Expression (2) by itself would lead one to try and adjust the circuit associated with the scintillation counter so that it reacts already on pulse sizes corresponding to the emission of only one photo-electron. As we will see presently, this is justified only if other fluctuations may be neglected.

From (2) we also see that in order to minimize σ_p, the product of m—in which the luminescent efficiency of the scintillation enters—and $1/\tau$ should be as large as possible. This makes stilbene better suited to fast work than anthracene; an efficient plastic scintillator should be still better.

Turning now to time fluctuations in the photomultiplier we will designate as σ_i the r.m.s.-deviation of the time in which a single electron traverses the i–th stage of the multiplier (stage 1 is photo-cathode to 1st dynode, stage 2 1st to 2nd dynode etc.). The *average* time for n_i electrons to traverse stage i will then have r.m.s. deviation $\sigma_i/\sqrt{n_i}$ so that the r.m.s.-deviation of the total transit time through the multiplier will be

$$\sigma_t = (\sum \sigma_i^2/n_i)^{1/2}\ . \qquad (3)$$

Now let the circuit following the detector react as soon as the output pulse reaches an amplitude corresponding to an average of n_1 primary photo-electrons. If a be the amplification per stage, assumed equal for all stages and if $\sigma_2 = \sigma_3 \cdots = \sigma_n$ we may then write

$$\sigma_t^2 \approx [\sigma_1^2 + \sigma_2^2/(a-1)]/n_1 = \sigma_m^2/n_1\ . \qquad (4)$$

[4]) R. F. Post and L. I. Schiff, Phys. Rev. **80** (1950) 1113.

The combined effect of scintillator and photo-multiplier is then given by the variance

$$\sigma^2 \approx \tau^2 n_1/m^2 + \sigma_m^2/n_1\ . \qquad (5)$$

This quantitiy can be minimized by taking $n_1 = \sigma_m m/\tau$, the minimum being

$$\sigma_{\min}^2 \approx 2\tau\,\sigma_m/m\ . \qquad (6)$$

If σ_m were small compared with τ/m, this minimum would have no physical meaning since then n_1 would have to be much smaller than 1. In that case, $\sigma = \tau/m$ is the smallest realizeable value, as already shown by (2). In practice, $\sigma_m \gg \tau/m$ in most cases. Often, n_1 can be chosen so as to minimize (5). We have already remarked, however, that σ_p increases more rapidly with n_1 than suggested by (2) as soon as no longer $n \ll m$. A larger minimum than (6) is then found for $\sigma_{\min}$, reached for $n_1 < \sigma_m\, m/\tau$.

Minton[5]), to whom I owe much of the argument presented here, has measured a value $\sigma_m = 1.8 \times 10^{-9}$ s for an RCA type 5819 multiplier operated at an overall voltage of 1 600 V. This spread appears to be caused almost entirely in the photo-cathode to first dynode gap (across which 180 V was used). In a scintillation counter consisting of a 5819 with a stilbene crystal we would have to choose $n_1/m \approx 0.3$ to achieve $\sigma_{\min} \approx 2 \times 10^{-10}/\sqrt{E}$ (E in MeV). This follows from (6) assuming that 500 photo-electrons are produced per dissipated MeV in the crystal and taking $\tau = 6 \times 10^{-9}$ s for stilbene.

Special efforts have been made lately at RCA to minimize σ_m, and it has been reported that their latest photomultiplier, type 6810A, which has a concave photo-cathode, shows a marked reduction in the transit time difference of photo-electrons starting from different points on the photo-cathode as compared with the type 6810. The transit time spread of the 6810A may thus even approach that of the good old 1P21.

It should be remarked here that a time spread of about 1 ns in the photo-cathode—1st dynode stage at normal voltages remains in all

[5]) G. H. Minton, J. Res. Nat. Bur. Stand. **57** (1956) 119.

photomultipliers that have the photo-cathode at the inside of the glass, even when only a very small part of the photo-cathode is illuminated. This spread can not be explained just as an effect of initial velocities of the photo-electrons, these would cause much smaller fluctuations. As far as I know, no satisfactory explanation has been offered, perhaps we must think of straggling of photo-electrons in the cathode and/or secondaries in the 1st dynode.

Minton observes that in a 5819 σ_m is inversely proportional to the square root of the voltage across the photo-cathode to dynode gap. This suggests the use of high voltages across this gap; unfortunately breakdown phenomena may often prevent us from increasing this voltage as much as we would like. The use of high *overall* voltages is attractive too from other points of view: the amplification increases rapidly with voltage and space charge effects become less serious, resulting in a better linearity. With existing multipliers one should, however, beware of breakdown and fatigue, when voltages much in excess of the rated values are applied continuously. In pulsed operation, on the other hand, considerably higher overall voltages may be used without ill effects. Singer *et al.*[6]) report operating 931 A's satisfactorily at total voltages up to 6000 V, pulsing them on for 0.1 μs. Peak currents of the output pulses up to 15 A could be obtained and these pulses showed rise-times of about 1 ns. Thus, for extreme time resolution, pulsed operation of existing scintillation counters, for instance in combination with pulsed sources, appears very attractive. On the other hand, we should also urge tube manufacturers to try and increase continuously permissible voltages in photomultiplier tubes.

A few words now about fluctuations in the photon collection time, neglected thus far. If large scintillators are used, as in cosmic ray research, or if long light guides are necessary between scintillator and photo-cathode, a noticeable spread in the time of arrival of photons at the cathode will occur. Just as in the case of the finite decay time of luminescence in scintillation crystals we can again derive an expression for the r.m.s.-deviation σ_1 of the moment of production of the n–th photo-electron. For a simple light guide and if $n \ll m$ (m is again the total number of photo-electrons produced per scintillation) σ_1 is given by an expression obtained from (2) simply by replacing the decay time τ by the minimum time t_0 it takes a photon to travel from the spot where it is created to the photo-cathode ($\sigma_1 \sim t_0\sqrt{n}/m$). This means that fluctuations in photon collection time become comparable to those in photon production time in fast organic scintillators for light guide dimensions of the order of a foot.

[6]) S. Singer, L. K. Neher and R. A. Ruehle, Rev. Sci. Instr. 27 (1956) 40.

3. Measurement of Nanosecond Time Intervals

Methods for measuring short time intervals between pulses from nuclear radiation detectors or between pulses from such detectors and a time reference pulse source find many applications in nuclear and meson physics. In the last few years some interesting new techniques have been developed for measuring time intervals in the nanosecond region. Effort has been directed at coding time difference information in such a way that it can be handled by multi-channel storage devices. A much used method of coding is time to pulse height conversion, in which case a multi-channel pulse height analyzer can be used to store the information. In the following, some methods for measuring nanosecond time intervals will be surveyed. We will start with

3.1. CONVENTIONAL FAST COINCIDENCE CIRCUITS

Some of the most used circuits of this type are actually Rossi-circuits, to which a very non-linear element, mostly a crystal diode, has been added to improve coincidence discrimination. A general requirement of the Rossi-circuit is that a number of current paths leading to a common point must be completely blocked. If vacuum tubes are used in these paths negative pulses of at least a few volts are needed to achieve this. A very familiar version of this

type of circuit is that introduced by Bell *et al.*[7]). In view of a modification of its operation to be discussed a little later I want to repeat here how it works. Negative pulses from two scintillation counters block limiter stages that deliver equalized positive pulses through delay cables to a common shorted shaping stub. To the input of this stub a crystal diode is connected, acting as a coincidence discriminator. This will conduct only if pulses from both channels overlap. The coincidences thus passed by the diode are then amplified and finally fed into a second discriminator, the setting of which determines the amount of overlap required to produce a coincidence count at the output of the unit. Time difference between counter signals is measured as the extra length of delay cable in one of the channels needed to bring the pulses in coincidence.

Some of the other conventional circuits in use for fast coincidence work have the advantage that they require no pulse equalization and will work satisfactorily with input pulses of only a few tenths of a volt. With the new photomultipliers, such as RCA's 6810A, pulse heights of a few volts are, however, easily realized, even for weak scintillations, Therefore, Rossi-type circuits such as Bell's and also Garwin's can be made to operate satisfactorily in most cases.

There may as yet be a special case for the so called

3.2. DIFFERENTIAL COINCIDENCE CIRCUITS

as introduced by Bay[8]) and also used by Minton[5]). The principle of differential coincidence operation is as follows: in conventional coincidence circuits there is an overlap region of the pulses at the coincidence mixer proportional to their widths. Since coincidences may also be passed when the overlap is not complete the coincidence curve will show a broadening dependent on pulse width and second discriminator setting. In a differential coincidence circuit events with incomplete overlap are rejected so that the width of the coincidence curve becomes almost exclusively determined by delay fluctuations in the multiplier.

It seems that *in principle* differential operation does not offer special advantages for obtaining ultimate time resolution, since also with a conventional circuit, e.g. Bell's, the width of the coincidence curve can be narrowed by using high first and/or second discriminator bias, and it appears that comparable widths are obtained at comparable coincidence efficiencies. It is possible, however, that better stability can be obtained with a differential coincidence circuit than with a conventional one set at high discriminator bias.

We will pass now to methods embodying *time to pulse height conversion* (3.3–3.5).

3.3. ACCUMULATED CHARGE METHOD

A very straightforward method of this kind consists of switching on a constant current that charges a capacitance each time at the beginning of a time interval to be measured and switching it off again at the end. The voltage pulses thus developed across this capacitance (C) will have amplitudes proportional to the width of the time intervals to be measured, they are fed into a pulse height analyzer. A

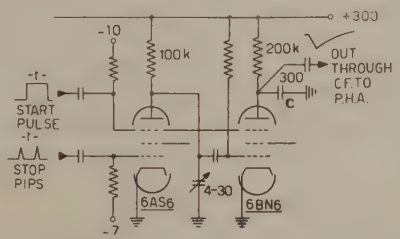

Fig. 1. Straightforward time to pulse height converter (ref. [8])).

circuit for measurements of time intervals in the nanosecond region in which this principle is employed, due to Weber, Johnstone and Cranberg[3]) is shown in fig. 1. It is used in a neutron time of flight spectrometer. The neutrons are produced in 2 ns bursts by a Van de Graaff generator in which a proton beam is

[7]) R. E. Bell, R. L. Graham and H. E. Petch, Can. J. Phys. **30** (1952) 35.

[8]) Z. Bay, Phys. Rev. **83** (1951) 242.

swept across a target. Pulses from the neutron detector are fed into a discriminator-pulse former which delivers the start pulses that initiate the charging of C by switching on the anode current of the 6BN6. These pulses have a width ($t = 270$ nsec) equal to the distance between neutron bursts. Stop pips, derived from the Van de Graaff beam deflection arrangement are fed to the first grid of the 6AS6 coincidence tube. The first pip that arrives after a start pulse will cause the 6BN6 to be cut off at its first grid. Capacitive stretching of the coincidence pulse at the 6AS6 anode prevents the 6BN6 from conducting until well after the start pulse has ended. Good time to height conversion linearity has been achieved except when stop pips arrive too near the edges of a start pulse. The authors state that the precision of time measurements is limited only by the characteristics of the detector. It seems, however, that these characteristics may affect the obtainable resolution rather seriously in this case as a result of time jitter caused by the amplitude spread of the neutron pulses in the discriminator pulse former.

3.4. OVERLAP TO CHARGE CONVERSION

A second, rather straightforward method of time to pulse height conversion has recently been mentioned by Sunyar[9]) (and it has apparently been used already some time ago in Chalk River). It uses Bell's coincidence circuit, just described. In this circuit, the charge passed by the discriminator diode when two pulses overlap at its input apparently is a measure for the degree of overlap, and therefore of the relative delay of the pulses. The output pulse height distribution of the rather slow amplifier that follows the diode reflects the charge distribution of the pulses passed through the diode and therefore also the delay distribution of the pulses from the detectors that fall within a time interval equal to the width of the shaped pulses. It is clear that really well equalized pulses are needed, since any size fluctuations will appear as delay fluctuations.

[9]) A. W. Sunyar, Bull. Am. Phys. Soc. 2 (1957) 137.

3.5. VERNIER METHODS

We now turn to the intriguing *vernier principle of time magnification*, introduced by Cottini, Gatti and Giannelli[10]). Since Dr. Gatti will say more about this I want to confine myself to mentioning the basic principle here and say a few words about statistics.

Pulses from two counters shock excite two *LC*-circuits tuned to slightly different frequencies (ω_1 and ω_2). Wave trains from these tuned circuits are mixed in a ring modulator, at the output of which the frequency difference $\omega_1 - \omega_2$ appears. Since the phase angle of the difference frequency is equal to the initial difference of the phase angles of the wave trains, a relative time shift of the wave trains by t seconds corresponds to a time shift $T = t\omega_1/(\omega_1 - \omega_2)$ of the difference frequency wave. Since T/t may easily be of the order 100 the shift T can readily be measured by a conventional time-to-height converting method.

As has been pointed out by Cottini and Gatti[11]) the phase of the shock excited wave trains is determined by the position of the centroid of the exciting pulses, provided these are short compared with the period of the wave. Therefore, the phase is independent of the amplitude of the pulses, at least as long as the circuit preceding the point where the shock waves are excited has a linear response. This is a very definite advantage of the vernier method.

The centroid of the pulses exhibits fluctuations in time as a result of finite phosphor decay time and transit time spread. We may approximate the variance of the total time spread by

$$\sigma^2 = \frac{\tau\sigma_m}{m}(\tau/\sigma_m + \sigma_m/\tau) \qquad (7)$$

which may be compared with eq. (6). It is seen from this comparison that σ approaches the value $\sigma_{\min}$ obtained in paragraph 2 if $\sigma_m \approx \tau$. This latter condition has indeed been realized by Cottini and Gatti, by using a scintillation phosphor with $\tau = 3 \times 10^{-9}$ s, they assume $\sigma_m = 2.10^{-9}$ s.

[10]) C. Cottini, E. Gatti and Giannelli, Nuovo Cimento 4 (1956) 156.

[11]) C. Cottini and E. Gatti, Nuovo Cimento 4 (1956) 1550.

The vernier method has also been adapted to time of flight neutron spectroscopy by Chase and Higinbotham[12]). In their instrument the wave-train from one of the detectors is replaced by a continuous wave derived from the r.f.-oscillator of a phase bunching cyclotron. This method has the advantage over the straightforward method mentioned under 3 of this paragraph that only a very small part of the r.f. cycle is lost for time analysis, even though the frequency is quite a bit higher (20 Mc/s in Chase and Highinbothams case).

A more literal version of a vernier method has been devised by Lefevre and Russell[13]). They describe a vernier chronotron, in which pulses from two detectors are placed on two circulating lines (fig. 2) of slightly different length. Each

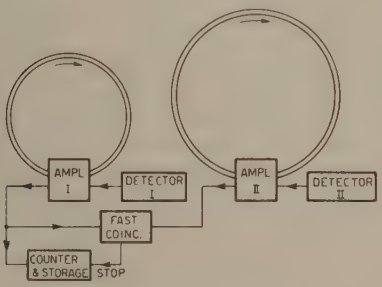

Fig. 2. Vernier chronotron (tentative block diagram, according to ref. [16])).

time after going round the line, the pulses are passed through saturating amplifiers. After a number of revolutions equal to the initial distance of the pulses divided by the delay difference of the lines they coincide. The coincidence pulse stops an address which had started counting when the first pulse arrived. Finally a pulse is added in the channel of the storage device at which the address had stopped. (Since only a cursory note in the Bulletin Am. Phys. Soc. about this instrument was available, part of the operation of this system is guessed.)

3.6. OSCILLOGRAPHIC METHODS

Time difference measuring devices should be mentioned in which use is made of an oscilloscope. The conventional oscillographic method is, of course, the photographic recording of properly triggered fast oscilloscope traces on which the pulses of which we wish to measure the distance appear. Because the whole detector pulse is displayed on the screen we can determine its position at least as accurately with this as with any of the other methods described so far. With modern travelling-wave oscilloscopes pulse intervals down to a few nanoseconds can be separated very well.

3.7. CHRONOTRONS

A last method to be briefly mentioned is the chronotron the principle of which is indicated in fig. 3. Pulses from two radiation detectors enter a fast line from opposite ends. Coincidence discriminators, distributed along this line, produce output pulses, the amplitude of which depends on the degree of overlap of the pulses at their inputs. These outputs are fed into a common slow line (also, a separate line for each detector can be used) and arrive at its right hand terminal with intervals of, for instance, 0.5 μs. This row of pulses can be displayed on an oscilloscope. From the pattern on the scope the relative delay of the pulses from the radiation detector can be determined. Ticho and Gauger[14]) describe a chronotron with a range of 20 ns and report a timing precision of $\pm$ 0.18 ns, using 5819 multipliers looking into liquid scintillators

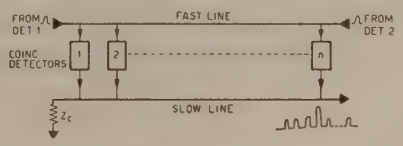

Fig. 3. Principle of chronotron.

traversed by cosmic rays. This precision indicates that the time resolution of the scintillation counters is the chief limiting factor also here.

[12]) R. L. Chase and W. A. Higinbotham, Rev. Sci. Instr. **28** (1957) 448.

[13]) H. W. Lefevre and J. T. Russell, Bull. Am. Phys. Soc. **2** (1957) 175.

[14]) H. K. Ticho and J. Gauger, Rev. Sci. Instr. **27** (1956) 354.

4. Conclusion

In almost any well designed instrument for measuring time intervals between pulses from scintillation- or Čerenkov-counters, the timing precision is limited by fluctuations in the detector. These are caused for an important part by transit-time spread in the photo-multiplier, as is revealed by a statistical analysis of the processes in the counter. It is to be hoped that photo-multiplier tubes with a further reduced transit-time spread will become available. At the moment their appears to be no fundamental reason why such tubes cannot be made.

DISCUSSION

GATTI pointed out that in the analysis it was the n–th electron which started the time measurement, whereas one really makes an average with different weights for the different photo-electrons.

DE WAARD agreed with this but said he wanted to get a lower limit.

BISHOP asked what was the minimum value of σ_1, and suggested that there were two possible contributions—one is the varying electron path lengths, while the other might be in the secondary emission process itself. DE WAARD thought this latter was normally considered small. There is also the possibility of straggling of photo electrons. Electrons can exist in the cathode for a finite time before combining. BISHOP pointed out that electrons penetrate to different depths according to the velocity and angle of approach, before they arrive at the surface and give rise to secondary electrons. This might give an ultimate limitation on σ_1.

VON DARDEL said that while Clyde Wiegand was at CERN he gave the latest information on the new R.C.A. photomultipliers. Apart from the 6810 A which has a curved inner surface, there is in development one with a curved inner and outer surface to reduce the straggling of electrons from different parts of the cathode to within 0.2 ns. The accelerating structure is the same, so σ_1 will remain the same. Dr. Morton at R.C.A. is attempting to put high potential grids between the dynodes to give a high field on the electrons when they come out of the dynodes, and this decreases the transit time through the linear structure by a factor of about 10. These improvements will give a factor 10 improvement on σ_1.

GATTI said that Dr. Bey of the Bureau of Standards suggested to him in a private communication just this idea of the grids, and in particular one very near to the cathode. It is not so very important to reduce straggling time in the later stages when there are many electrons, but it is important to have a low straggling time when the first electron is involved.

MOODY commented on the use of diodes in the Fast Coincidence Circuit and the assumption that the characteristics are linear with time—and pointed out that this is not true because the impedance of semi-conducted diodes is a function of time. These effects apparently do not entirely stop one from doing these things because he did them some seven or eight years ago, almost in ignorance of the effects. The characteristics of diodes are now well known and they have been measured—one should not neglect them when using diodes for fast pulse work.

DE WAARD asked if there had not been a considerable improvement in diode characteristics since then.

MOODY said it was difficult to give a quantitiative answer since there are so many parameters and diode types; the best turn off time is about a 10 to 1 fall in current in 0.03 μs. But it is easy to find diodes with 0.1 μs fall time and even higher. Turn on time will be of the order of 0.01 μs. Such good diodes usually are characterised by a high forward current at quite a low voltage, and an inability to withstand a reverse voltage of more than 10 volt.

NUCLEAR INSTRUMENTS 2 1958 (81–87); NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION I.2

SOME COMPONENTS USED IN THE NANOSECOND FIELD

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1. Introduction

In this paper I will try to tell some things that might be of interest to you, as users of components.

The first components we think of are *valves*.

In this special field, the general problem is that one wishes to be able to generate waveforms with the shortest possible rise and fall time. Let

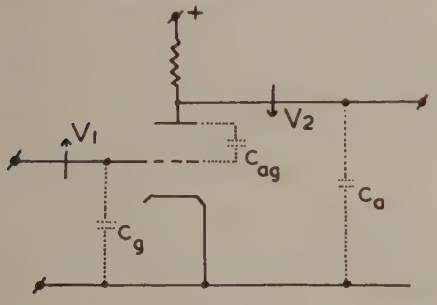


Fig. 1

us consider an unstable flip-flop; it consists in essence of two video stages in cascade, with inputs and outputs interconnected. The shortest attainable rise and fall times are given by the gain-band-width product, which gives us the highest frequency that can be generated.

In this simple circuit (fig. 1) the GB-product is proportional to $G_m/(C_a + C_g + 2C_{ag})$; $C_a + C_g + 2C_{ag}$ is the capacitive loading of the anode resistor $1/G_m$. The a.c. voltage on C_{ag} is twice that on the other capacities, so that $4C_{ag}$ has to be written in the energy balance. In the form (1) $2C_{ag}$ is included in C_g and C_a, so that in effect C_{ag} occurs 4 times.

For instance, in the "Bell System Technical Journal" (Oct. 1951) the various factors that limit broadband performance are studied. It is shown that the distance between cathode and grid has by far the greatest influence on the GB-product, and a decrease in distance will improve the performance of the valve.

How far can be gone in decreasing the distance depends on the mechanical design. Now, in modern valves stretched first grids are used. These grids contain a rigid frame, around which the fine grid wire is spiralled. In the finished grid wires there is a high residual tension that keeps them in the grid plane, even if the grid wire is much finer than has been used in conventional grids.

A few typical valves are listed in table 1.

TABLE 1

Tube type	V_a	V_{g_2}	I_a	G_m	C_a	C_g	$C_g+C_g+2C_{ag}=C_1$	$C_g+C_g+2C_{ag}+10=C_2$	$G_m/2\pi C_1$	$G_m/2\pi C_2$
EF95–6AK5	180	120	7.7	5.1	2.8	4.0	6.84	17	114 Mc	48 Mc
E180F–5A/170K–6688	190	160	13	16.5	3	7.5	11	21	240 Mc	126 Mc
435A		150	13	15	2.5	7.8	10.35	20	220 Mc	114 Mc
437A			40	45	4.4	15	26.4	37	280 Mc	194 Mc
E88CC	100		15	12.5	3.2	4.7	10.7	20	186 Mc	100 Mc
E88CC	100		30	25	6	6.8	18.4	28	210 Mc	142 Mc
EFP60			20	25	6	9.2	15.2	25	260 Mc	160 Mc
Z319			15	21	4.9	9.1	14	24	240 Mc	140 Mc

The GB-product, $G_m/2\pi C$, is given
(a) for the tube itself and
(b) for the tube with 10 pF wiring capacitance added.

The 6AK5 (EF95) is of conventional construction, and it is evident that with the following more advanced valves a remarkable improvement is obtained. The newest valve of this type is the E88CC, a double triode intended for use in cascode circuits and fast flip-flops.

The E88CC is also listed for the situation when one uses the two halves in parallel; an improvement is obtained in this case, as the internal screen need not be grounded, and the wiring capacitance is of less influence.

The valve is able to draw a relatively high current at low anode voltage, and will not lose its emission when kept for a long time in cut-off condition.

A special class of valves is formed by the EFP60 and the Z319 secondary emission tubes. Their transconductance is not higher than one encounters in other more modern valves, but they incorporate the feature of providing two output signals of opposite polarity, and they have a current gain between cathode and anode.

The worst drawback of these valves is that they have never attained the degree of reliability and constancy of other valves.

One might ask if a better secondary emission valve could be made. A high transconductance can be achieved in a secondary emission valve if the primary grid-cathode structure has a high value of G_m/I_a. Now this value increases with decreasing I_a; the maximum of about 10 mA/V/mA occurs at very low cathode current densities.

At this maximum the c.g. distance has no further influence, but at higher densities an improvement is obtained with decreasing distance.

It is theoretically possible to obtain still higher values of G_m/I_a, but only with other systems, e.g. with deflection tubes or double-beam systems.

Other valves that may be of interest are
(a) Valves with more than one grid with comparable cut-off caracteristics. An example is the EH91, which has a g_1 and g_3 that both will cut-off the anode current at about 10 volts, under normal conditions. It has been designed as a gating valve in computers, and is similar to the 6BN6.

(b) Deflection valves. A typical example is the E80T. This type of valve has small capacitances, and used as a gate valve there is a very good shielding between input and output when the signal must not pass the valve.

(c) The type of valve that is intended to be used at an anode voltage of 6.3 or 12.6 V in car radiosets.

A typical caracteristic, of a pentode used as triode, is shown in fig. 2. Such a valve can be

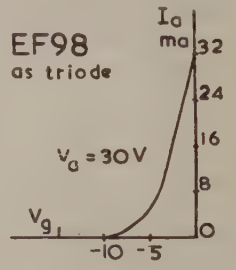


Fig. 2

useful when a high anode current at low anode voltage is needed.

2. Diodes

There is one type of thermionic diode that might be of interest in high-resolution circuits; the EA52.

This diode has a disc seal cathode, and an anode rod pointed towards it. The inpedance in the forward direction is about 1000 Ω at 10V; in the inverse direction higher than $10^4 M\Omega$.

The capacitance Ca is 0.5 pF, and the resonance frequency in the region of 4000 Mc.

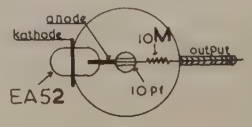


Fig. 3

This diode can detect very narrow pulses that travel along a cable without causing appreciable

reflection, and it seems therefore well suited for insertion in a chronotron circuit where a lot of diodes are connected to a short length of cable.

A typical example of the construction of a detector on a cable is shown in fig. 3.

3. Transistors

We can apply a reasoning of the type we used with valves also to the transistors.

Again we can consider a flip-flop to be a cascade of two video-amplifiers.

The circuit of one stage, and the simple equivalent circuit for the transistors, are given in fig. 4.

Fig. 4

In this equivalent circuit, the influence of the collector voltage on the emitter-base region is neglected. This is possible when the collector a.c. voltage is small, which indeed is the case when we consider a circuit with low output impedance.

The relaxation time RC is such that $R = \omega C$ at $f = f_a^{co}$; again this circuit gives a good approximation if the impedance of the signal source is low.

For such a simple video stage the gain-bandwidth will be nearly equal to f_a^{co}.

With valves it is easy to trade band-width for gain and vice versa. With transistors however the situation is more complicated, but the gain bandwidth product can be maintained up to $\frac{1}{2}f_a^{co}$ when a suitable feedback is applied.

With the most promising transistors, the drift (graded base) transistor, we can now achieve a f_a^{co} of 600 Mc/s, which brings the best transistors into the region of good valves. So, under favourable conditions, we may expect to obtain a risetime of the order of 10^{-9}–10^{-8} s.

In table 2 there are listed a few different types of transistors. The grown junction, the alloy junction and the diffused junction transistors represent one class of transistors, which differ in properties mainly because they were prepared by different methods.

In the grown junction type the different regions are produced by adding impurities to the melt during formation of the crystal.

The alloy junctions are prepared by alloying the already grown crystal with an impurity containing metal.

In the drift transistors, the impurity material is diffused from the vapor phase, and it is possible in this case to obtain highly plane-parallel junctions with a special impurity gradient. The charge carriers can be accelerated and base transit time reduced, which results in a high f_a^{co} combined with low base spreading resistance and low collectorbase capacitance.

The surface-barrier transistor is made using an electroplating technique. This method allows for a very thin base region; the result is a high f_a^{co} and a low collector power rating.

The double base transistor has two connections

TABLE 2

	Cut-off freq.	Dissipation	Bias (V_c)	Input (Ω)
Point contact triode	1–> 100 Mc	100–400 mW	30	0–200
NPN grown junction	1–20 Mc	20–50 mW	5	50–100
PNP alloy junction	0.5–> 20 Mc	20 mW–30 W	5–40	50–100
PNP diffused junction (drift)	1–600 Mc	30–50 mW	10	50
NPN double base junction	10–100 Mc	20–50 mW	5	50–100
Surface barrier	30–> 100 Mc	2–10 mW	3	30

to the base region. A current is passed between the two connections, which reduces the effective area of the base greatly. This also has the effect to improve f_α^{co} and to lower the base spreading resistance.

4. Crystal Diodes

The behaviour of crystal diodes when a rectangular voltage pulse is applied, is given in fig. 5.

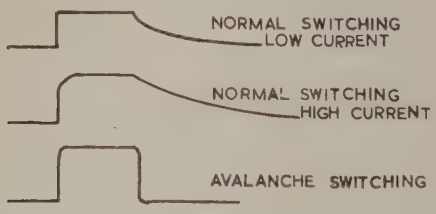


Fig. 5

In the case of small voltages the switching time from zero to forward current is very small; at the end of the pulse however there is a tail in the current pulse due to minority carrier storage effect.

The relaxation time of this effect is in the order of microseconds. At high currents there is also a relaxation effect at the start of the pulse, due to conductivity modulation, in the order of a fraction of a microsecond.

Now Scobey, White and Salzberg (Proc. IRE (Dec. 1956) 1880–1881) report that a very small switching time from high to low currents can be achieved by using the avalanche region of a junction diode; Salzberg and Sard (Proc. IRE (Aug. 1957) 1149–1150) state that an improvement is also obtained for transitions from low to high currents.

They report that with "currently available equipment" (Hewlett Packard type 212A pulse generator, and Tektronix type 545 oscillosope with type 53–54 K preamplifier) the transients are so short as to challenge measurement.

5. Transformers

Whether we can use transformers in the nanosecond region depends on the requirements to be fulfilled.

It can be derived that for a simple two-layer winding the maximum frequency at which the transformer still works as a transformer is determined by the length of the wire of the primary or secondary coil, whichever is longer.

For instance a transformer with a wire length of 5 cm will exhibit a resonance frequency of about 10^9 c/s.

Now if we wish to have impedance transformation we will have to put at least some turns on the coil with the longest wire; let us choose, 5 turns on a core of 3 mm.

An insulation of 0.1 mm between primary and secondary leads then to a ωL leakage of about 60Ω at 10^9 c/s, and a primary self induction of about 1000Ω, assuming a μ of 30 in the core.

The ΔB of this transformer will be $\approx$ 500 gauss at short pulses of 100 V; longer pulses of small amplitude (duration e.g. 0.5 μs) can be transmitted with a much smaller distortion than one should expect, because the μ of the core is dependent on the magnitude of the pulse, as will be discussed later.

A transformer of the type described here is by no means ideal; it will not handle a high recurrence frequency, and both high and low frequencies suffer distortion.

The transformer can give us phase reversal and impedance transformation; if we drop one of the two, the frequency response can be improved greatly.

First, if we drop the impedance transformation, the structures of figs. 6(a) and (b) can be made.

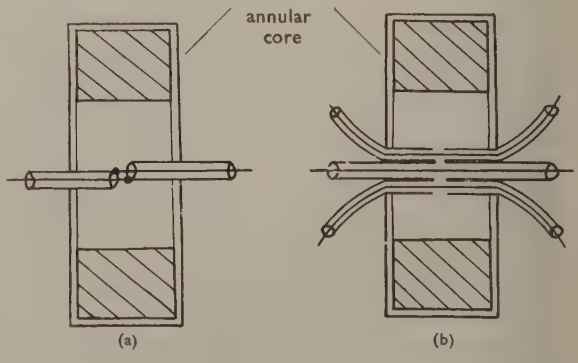


Fig. 6

In fig. 6(a) a device is shown that inverts the polarity of signals up to very high frequencies; the high frequency response is essentially that of a cable, and the low frequency response depends on the amount of ferromagnetic material we use.

Such a device, used in combination with one of the new highslope triodes or tetrodes, might be a good substitution for a secondary emission tube.

In fig. 6(b) a similar construction, but with 3 feedthrough cables, is shown. This transformer has been used in the differential anticoïncidence circuit of fig. 7.

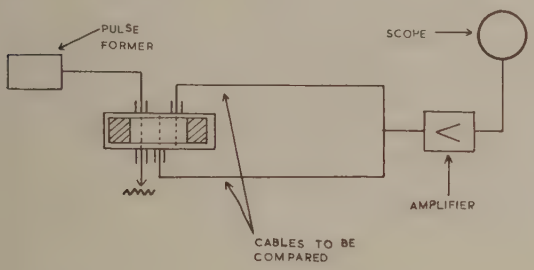


Fig. 7

In this circuit a pulse of about 300 V, 10^{-8} s, 50Ω is split into two pulses of opposite polarity and fed through two identical cables.

The output pulses are added; the sum of them (very small compared to the pulses themselves) is amplified and examined with an oscilloscope. When the two signals are of equal shape and magnitude, and the distortions in the cables are equal, a very small difference in transit time in the two cables can be measured.

The difference between the signals from the transformer was here about 1/3000 of the signals themselves; with short cables (which means small difference between the cables) the signal on the scope about doubled when a transit time difference of 3×10^{-12} s was introduced.

This high resolution was only possible because we compared here two identical signals. When the signals have not the same shape it is always difficult to measure a time interval between them of a duration which is short compared with the pulse; indeed even the definition of the time difference itself is not clear. With longer cables (1 m–10 m of RG 58–U) the resolution obtainable was worse, as the distortion was apparently too different for the two cables.

With cables of 10 m a difference of 5 mm was clearly visible.

6. Tapered line

Another type of transformer is the tapered line. This device can give impedance transformation. The high frequency performance is that of a line, but to achieve good low frequency response, the tapered line has to be made very long.

This device is extensively dealt with by Lewis and Wells (Millimicrosecond Pulse Techniques), who gave the formula.

$$\frac{\text{Total delay along line}}{\text{pulse duration}} = \frac{50\,(\ln N)^2}{P}$$

where

N = voltage transformation ratio

P = percentage of fall of amplitude at the end of the pulse.

Of course a long line will give us a long transit time and some attenuation of the higher frequencies.

7. The Core

Now let us look for a moment at the properties of transformer cores in this region.

If we apply a rectangular voltage pulse to a transformer, the magnetic induction will increase linearly with time.

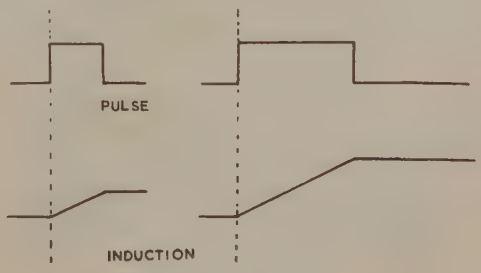


Fig. 8

For a short and long pulse fig. 8 shows some curves, for the case of voltage pulses of equal magnitude.

In the beginning of a long pulse the situation in the core is exactly the same as during the short pulse; the core has no knowledge of the future and will therefore have the same permeability in both cases.

Apparently the μ will be dependent only upon dB/dt, so as to give a low μ at a high dB/dt and vice versa.

Now it is difficult to predict the behaviour of a core under pulse conditions from the data obtained with sinusoidal varaitions.

The best thing one can do is to measure the B–H loop of the material under pulse conditions. As far as nickel iron cores are concerned, a lot of data is available from the work of W. S. Melville.

For ferrite cores, there is little data available, and I will end with a description of some measurements. The intention was to make visible the B–H loop, under pulsed conditions, of a small ferrite ring (diameter 3 mm, cross section 1 mm^2).

A rectangular pulse current was applied to the ring; rise and fall times were of the order of 0.02 μs, and pulse duration 0.1 μs.

As a pulse source we used a line-type modulator with the hydrogen thyratron 4C35, as a compromise that enabled us (1) to get a high pulse voltage compared to the induced voltage in the ring, (2) to keep the pulse source small and well screened to minimise error voltages in the measuring circuit.

The magnetic field in the ring was approx. 30 oersted, the signals for E and I were a few volts. The schematic diagram is shown in fig. 9.

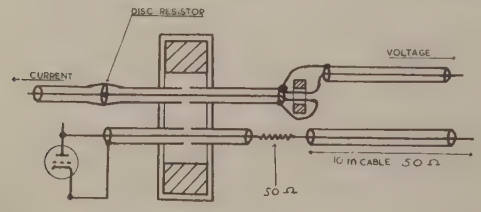


Fig. 9

In this diagram the transformer was used to inject the power in the measuring circuit in a simple way; it also provides d.c. isolation and permits easy reversal of the pulse on the ferrite ring.

Provision was made for a d.c. current in the magnetizing winding. The voltage coil consisted of one turn, the current coil of 3 turns; with the pulse circuit used the induction in the ring could be inverted within 10^{-8} s from one saturation direction to the other.

To obtain the B signal it is necessary to integrate the E signal. First a RC integrator was tried. It is difficult then to conduct the signal from integrator to oscilloscope. Worse is that the integrated signal is so small as to make it practically impossible to keep unwanted signals away from the picture.

An LR integrator, employing a cable as R, solves the first problem; the unwanted signals where the same as with the RC integrator however.

Signals resulting from imperfect shielding of the cables were always present. In our case the signals E and I were fed to mixing circuits of a sampling-type oscilloscope. As in this type of oscilloscope there is a signal present that is a low frequency analogue of the observed high frequency signal, it was possible to integrate this low freq. signal in a low freq. integrator, thus eliminating the use of a high freq. integrator altogether.

Finally the following signals were displayed:

(1) H against time.

(2) B against time.

(3) H against B.

The 3 curves were shown in 3 colours on a coulour television monitor, adopted for use with the sampling scope.

Examination gave for transitions from $-B_{max}$ to $+B_{max}$ in 1/100 μs a permeability of 30, for slower transitions a permeability rapidly increasing to about 1000. Of course the length of the cables between sample and mixer, transmitting the E and I signal, has to be carefully matched.

A difference of 25 cm gave a marked difference in width of the B–H loop.

When cables of equal length were used, a small coil without ferrite, and of about the same inductance as the measured coil, gave as B–H loop a straight line, as might be expected.

DISCUSSION

SCARROTT said that ferrite has a much lower permeability if you switch it fast—could you make the calculations easier if you represent this by a shunt resistance? If you assume that the mutual inductance is an inductance with a shunt resistance, you will get this effect.

MOODY said that in switching-cores the phenomenon of domain relaxation time is rather significant. It seems likely that the less hard ferrites might also have such a time. For the hard ferrites it is necessary to wait a little time before it is evident.

BRUIJNING replied that it is difficult to compute behaviour with fast transients when only small signal data is available.

NUCLEAR INSTRUMENTS 2 (1958) 88–89; NORTH-HOLLAND PUBLISHING CO., AMSTERDAM

SESSION I.3

IMPROVEMENT IN THE VERNIER TYPE NANOSECOND TIME DELAY ANALYZER

C. COTTINI AND E. GATTI

C.I.S.E. and Politecnico, Milano

As a result of a stimulating discussion with Dr. Fidecaro (CERN) we are improving our Vernier type analyzer[1,2,3]) to make it suited to higher counting rates up to 10^5 single pulses per channel. The new instrument is not yet completed but partial results on the modified parts indicate that no worsening of the previous results is expected.

The main changes are the damping of the resonant circuits made by a symmetrical couple of normally conducting diodes as shown in fig. 1.

Damping is introduced in order to prevent the use of dead time in both channels, which otherwise should be provided in order to remove errors due to superposition of excitation of the resonant circuits due to singles.

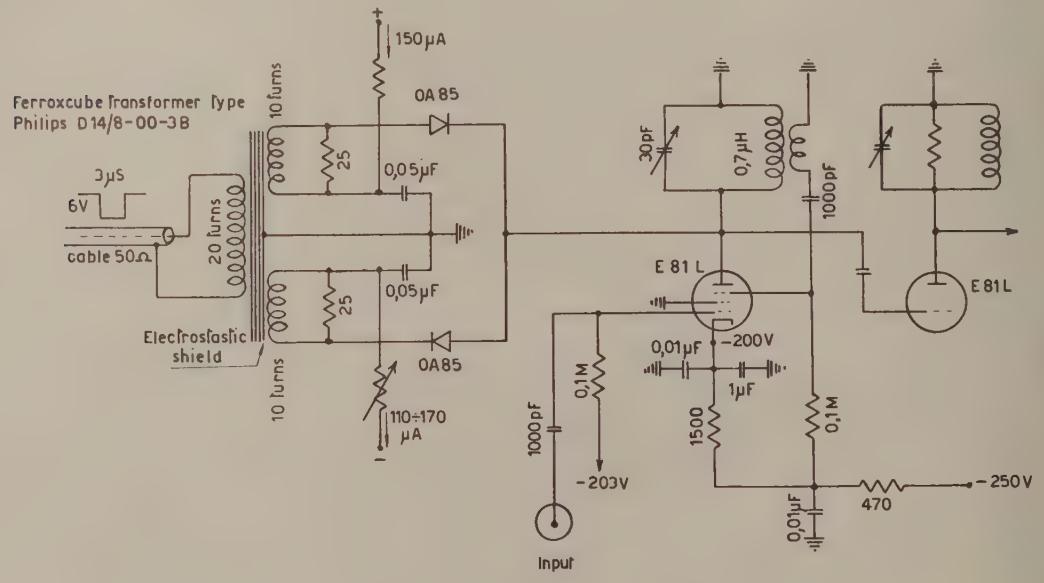


Fig. 1

The quiescent current fine regulation of the damping diodes allows nearly perfect balancing in order not to excite the resonant circuits when damping is removed.

Residual excitation is about 1/50 of the one due to the smallest pulses accepted at the input (0.1 V) and this worst case corresponds to a time spread of 1.5×10^{-10} sec.

Further the losses in the first resonant circuit are compensated by a slight positive feedback applied through the suppressor grid as also shown in fig. 1.

[1]) C. Cottini, E. Gatti and G. Giannelli Nuovo Cimento 4 (1956) 59.

[2]) C. Cottini, E. Gatti, Nuovo Cimento 4 (1956) 1550.

[3]) S. Colombo, E. Gatti and M. Pignanelli, Nuovo Cimento 5 (1957) 1739.

This is done in order to make an easier limitation of the sinusoidal waveform by the 6BN6 tube.

following a quasi-coincidence by paralysing the phototube by inversion of the voltage between two adjacent dynodes.

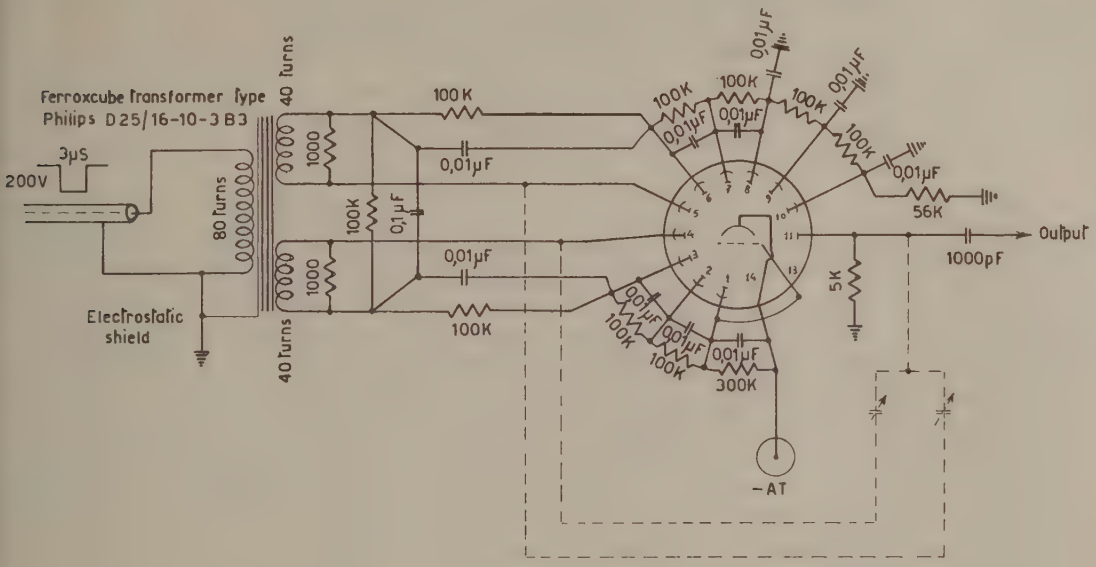


Fig. 2

"Linear" and "non linear" way of working is provided[3]).

When working with the linear method, the measurements are protected against pulses

Here again spurious signals are completely eliminated by a capacitive bridge balance as shown in fig. 2.

NUCLEAR INSTRUMENTS 2 (1958) 90; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION I.4

ANALYSEUR DE TEMPS DE VOL DANS LE DOMAINE DES NANOSECONDES; SOMMAIRE†

R. MEUNIER

Centre d'Etudes Nucléaires de Saclay, Gif-sur-Yvette (S & O)

Un analyseur de temps de vol a été construit sur le principe du chronotron, en lui adjoignant un interpolateur électronique qui définit, de façon continue, le retard relatif de deux pulses provenant de compteurs à scintillation.

Ce délai est défini par les "centres de gravité" des deux pulses. Le résultat se présente sous la forme de la déflection radiale d'un spot cathodique dans une direction formant un angle proportionnel au retard à mesurer. Les tensions nécessaires à cette déflection sont obtenues par addition vectorielle des douze pulses fournis par les douze détecteurs du chronotron. Le retard mesuré ne dépend pas de la hauteur des pulses fournis pas les compteurs. Le temps de résolution est défini par les pulses des compteurs. Il est environ 3 à 10 fois plus petit que la durée totale d'un pulse et des temps de résolution de 5×10^{-10} s ont été obtenus avec des compteurs. Il n'y a aucun délai dans la mesure d'un retard et aucun circuit de coïncidence lent n'est nécessaire. Les mesures de retard peuvent être effectuées jusqu'à des cadences d'un million par seconde.

† Un article sur ce même sujet a été publiée à la Rev. Sci. Instr. 28 (1957) 1010–1015.

DISCUSSION

MEUNIER informed that he did not in his circuit use any slow coincidence circuits. The spot must of course be blanked when it is in the centre of the screen to reduce the unwanted light. Resolution is given by total time to do computation—this is less than 1 μs, much longer than the 30 ns it takes the pulse to go along the line.—The theoretical limitation is given by the length of the line.

To a question by HUTCHINSON whether the diodes remain matched MEUNIER replied that after one month of continuous use there was no drift. But there may be changes due to temperature.

The best diodes were found to be Transitron S6 Silicon diode, and the French type DPO made by S.F.O.

NUCLEAR INSTRUMENTS 2 (1958) 91–94; NORTH-HOLLAND PUBLISHING CO. · AMSTERDAM

SESSION I.5

RÉALISATION DE CIRCUITS ÉLECTRONIQUES RAPIDES DANS LE CADRE DU STANDARD C.E.A.

J. MEY

Centre d'Etudes Nucleaires de Saclay, Gif-sur-Yvette (S & O)

1. Introduction

Il y a environ un an et demi les physiciens qui avaient en vue l'utilisation du synchrotron à protons dont la construction s'achève à Saclay, firent appel au Service des Constructions Electriques. Pour leurs travaux futurs ils prévoyaient le besoin de quantités importantes de matériel électronique spécialement adapté aux expériences de physique des hautes énergies c'est-à-dire principalement de circuits rapides. Devant l'importance des travaux il fut décidé de les répartir entre plusieurs participants. D'une part le Service des Constructions Electriques aidé par l'industrie privée et d'autres laboratoires d'étude s'occuperait des appareils destinés à une certaine diffusion, d'autre part le groupe de physique des hautes énergies étudierait les problèmes plus particuliers et devant conduire à la réalisation d'un nombre restreint d'exemplaires. Toutes ces études et réalisations devaient cependant être conduites dans un certain esprit de standardisation et aboutir chaque fois que cela était possible à la création d'éléments fonctionnels interchangeables. Pour assurer une certaine continuité et accroître les possibilités d'utilisation, il fut décidé d'utiliser les mêmes éléments mécaniques interchangeables que pour l'électronique plus classique, éléments qui sont actuellement employés couramment au Commissariat à l'Energie Atomique. Les nouvelles réalisations en électronique rapide devaient ainsi constituer une extension de l'électronique standard. Par exemple une échelle de comptage rapide devait utiliser pour les derniers étages les éléments de l'échelle classique tandis que les premiers étages seraient constitués d'éléments interchangeables nouveaux plus rapides.

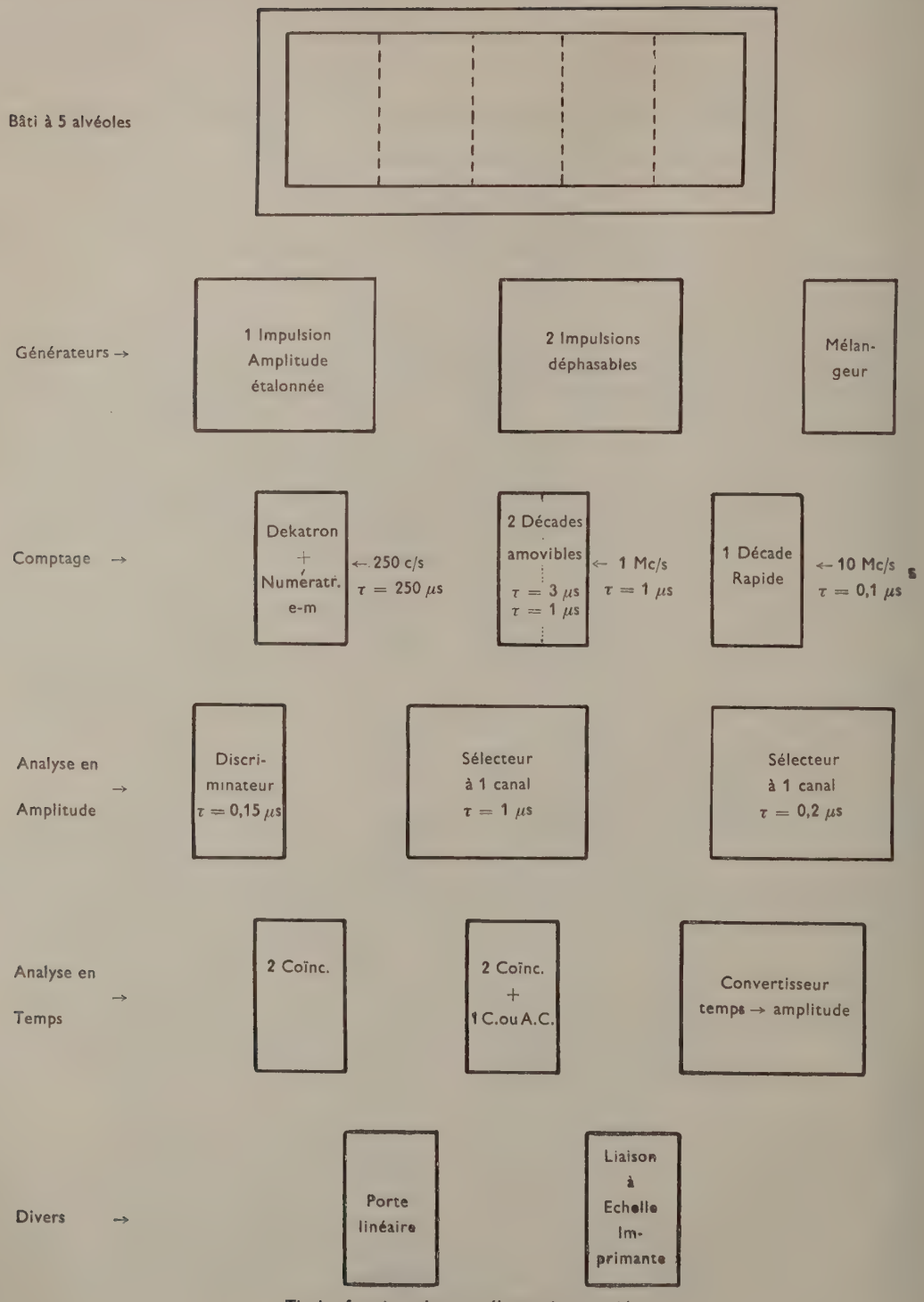
Le standard mécanique actuellement adopté au C.E.A. est basé sur l'utilisation de bâtis comportant cinq alvéoles dans lesquels peuvent être logés des tiroirs amovibles. Ces tiroirs ont une hauteur de 175 mm, une profondeur utile de 263 mm et existent en trois largeurs différentes pouvant occuper respectivement un, deux ou trois alvéoles, l'unité correspondant à 78.5 mm. A l'arrière des bâtis peuvent être placées des sources d'alimentation et de polarisation. Une plaquette d'interconnexions permet d'y effectuer les différentes liaisons électriques entre ces sources et les tiroirs ou entre les tiroirs eux-mêmes. Ces liaisons se font par l'intermédiaire de fiches à 14 broches.

2. Eléments fonctionnels d'électronique rapide

Un certain nombre de tiroirs fonctionnels ont été réalisés depuis. Ils sont destinés à remplir les différentes fonctions demandées dans les expériences, les principales étant le comptage des impulsions, leur analyse en amplitude, leur analyse en temps. De même pour effectuer les essais de ces circuits des générateurs d'impulsions ont été construits. Nous allons décrire rapidement ces diverses réalisations.

2.1. GÉNÉRATEURS D'IMPULSIONS

Deux modèles ont été réalisés, chacun dans un tiroir deux unités. Le premier fournit une impulsion de forme rectangulaire étalonnée en amplitude. Il utilise un relais à contacts mouillés au mercure et une mise en forme par ligne coaxiale ouverte à une extrémité. Le



Tiroirs fonctionnels pour électronique rapide

relais (Western Electric ou Clare) a été sorti de son enveloppe métallique et introduit dans une structure coaxiale afin d'assurer une bonne adaptation des impédances. L'amplitude des impulsions peut être réglée à l'aide d'un potentiomètre hélicoïdal à 10 tours, de 0 à 40 volts. La durée du signal dépend de la ligne de mise en forme choisie. La fréquence de répétition définie par un oscillateur incorporé est au maximum de 40 c/s. L'utilisation de la fréquence du secteur a été écartée d'une part pour éviter tout couplage des circuits par l'intermédiaire du secteur et d'autre part pour avoir une tension d'excitation du relais et par suite une résistance de contact bien constantes. Le second modèle produit deux impulsions de largeur à la base 30 ns dont on peut régler le déphasage d'une façon continue. Les amplitudes peuvent être réglées séparément. Le générateur est déclenché par un oscillateur extérieur ou par le signal dent de scie d'un oscilloscope d'observation, jusqu'à des fréquences de 50 kc/s.

Pour la mesure du temps de résolution des circuits de comptage il est souvent utile de disposer de quatre impulsions déphasables. Ceci est possible par l'utilisation du générateur précédent et d'un circuit de mélange construit dans un tiroir simple. L'une des entrées de ce circuit est alimentée directement par le générateur de paires d'impulsions tandis que l'autre est alimentée par ce même générateur mais à travers une boîte à retards règlables.

2.2. FONCTION COMPTAGE

Pour cette fonction l'utilisateur dispose de trois types de tiroirs simples lui permettant de réaliser par leur assemblage judicieux une échelle ayant la capacité et le temps de résolution voulus. Le premier est emprunté à l'électronique standard. Il comprend un tube décade à gaz (Dekatron) suivi d'un numérateur électromécanique à 4 chiffres et à remise à zéro manuelle. Ce dernier pouvant compter jusqu'à 25 coups par seconde, le tiroir admet à son entrée des impulsions récurrentes à la fréquence maximum de 250 c/s, son temps de résolution en statistique étant de 250 μs. Le second type est semblable au tiroir double décade standard. Celui-ci est constitué de deux décades identiques câblées sur plaquettes amovibles et ayant un temps de résolution de 3 μs. Une nouvelle décade plus rapide, ayant un temps de résolution de 1 μs, a été réalisée sous la même forme et peut être mise à la place des plaquettes standard. Le troisième type contient une décade rapide de temps de résolution 0.1 μs. Enfin un tiroir encore plus rapide est à l'étude et est destiné à précéder tout cet ensemble.

2.3. FONCTION ANALYSE EN AMPLITUDE

Plusieurs modèles de tiroirs discriminateurs ont été réalisés avec des temps résolution compris entre 0.1 et 0.2 μs et une sensibilité variant de 0.25 V à 1 V. Le modèle le plus courant est un tiroir 1 unité possédant un temps de résolution de 0.15 μs et un seuil réglable de 0.5 V à 20 V par potentiomètre hélicoïdal 10 tours. La dérive du seuil, avec chauffage stabilisé à 1% est de 5/100^{e} volt sur 10 heures avec des fluctuations maximum de 2/100^{e} volt par demi-heure. Une version améliorée va en être réalisée avec entrée en positif ou négatif sorties en positif et en négatif par l'intermédiaire d'un circuit porte.

Deux types de sélecteurs d'amplitude à 1 canal ont été construits dans des tiroirs doubles. L'un utilise les méthodes classiques d'anticoïncidence et a un temps de résolution de 1 μs. L'autre, d'un type spécial possède un temps de résolution de 0.2 μs. La bande d'analyse est définie par les seuils de deux discriminateurs. L'impulsion à analyser est envoyée directement sur le discriminateur de seuil supérieur et à travers une ligne à retard sur celui d'inférieur. Si l'impulsion est comprise dans la bande ce dernier déclenche seul et fournit un signal de comptage. Par contre si son amplitude est plus grande que le seuil supérieur le discriminateur attaqué directement déclenche le premier et produit un signal d'interdiction qui est envoyé sur l'autre discriminateur pour en empêcher le déclenchement lorsque l'impulsion, retardée dans la ligne, se présente à son entrée.

2.4. FONCTION ANALYSE EN TEMPS

Deux modèles de circuits de coïncidence ont été réalisés dans des tiroirs simples. Ils sont tous deux du type Garwin et utilisent des pentodes à grande pente E 180 F. Le premier possède deux voies de coïncidence en négatif avec une sensibilité de 1 volt. Le temps de résolution est de l'ordre de 2 ns et dépend de la mise en forme des impulsions qui s'effectue par câble coaxial court-circuité placé sur les grilles d'entrée. Le second, de performances identiques possède trois voies de coïncidence en négatif dont une utilisable en anticoïncidence en positif grâce à un commutateur.

Pour la sélection en temps un convertisseur temps-amplitude, construit dans un tiroir double, permet étant associé à un sélecteur d'amplitude multicanaux de réaliser un analyseur en temps. Le rapport de conversion est de 0.7 V par ns.

2.5. FONCTIONS DIVERSES

Enfin plusieurs tiroirs ont été construits pour remplir des fonctions diverses. En particulier un tiroir porte linéaire destiné à être placé sur une voie d'analyse et à assurer la transmission ou l'arrêt des impulsions sous l'action d'un signal de commande. Dans le cas de la transmission, la linéarité est de 1% pour des amplitudes allant jusqu'à 20 volts. Egalement un tiroir de liaison à une échelle imprimante. Placé à la sortie d'une échelle de comptage purement électronique, son rôle est d'effectuer sur cette échelle un certain nombre d'opérations en liaison avec une échelle de lecture équipée pour l'impression des résultats. Il commande l'arrêt de l'échelle, le transfert du résultat affiché sur l'échelle imprimante, la remise à zéro puis en état "marche" de l'échelle de comptage.

3. Conclusion

En fait il n'est pas possible de réaliser tous les circuits rapides ou auxiliaires sous forme d'éléments fonctionnels facilement juxtaposables. Certains circuits soit plus délicats, soit plus importants nécessitent une réalisation globale. Néanmoins l'emploi de tiroirs fonctionnels interchangeables, chaque fois que celà est possible, doit permettre à l'utilisateur de composer rapidement et à volonté un ensemble de mesure. Cette souplesse de construction autorise le changement immédiat d'une partie de la chaîne, soit pour obtenir des performances différentes, soit simplement pour en vérifier le bon fonctionnement ou éventuellement remplacer un élément défectueux.

NUCLEAR INSTRUMENTS 2 (1958) 95; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION I.6

RECENT WORK ON NANOSECOND TECHNIQUES AT HARWELL

P. R. ORMAN

Atomic Energy Research Establishment, Harwell, Didcot, Berks

Summary. The author gave details of four instruments which have been developed recently at Harwell.

(1) Distributed Amplifier, with rise-time of 2.5 ns and gain of 20 db with printed wiring. See Nuclear Instruments **1** (1957) 354.

(2) Nanosecond Pulse Generator. See Nuclear Instruments **1** (1957) 183. Orman and Wells.

(3) Time-Amplitude Converter for use in the nanosecond region. Time is measured between a counter pulse and the next marker pulse. Marker pulses arrive with a repetition frequency of about 10 Mc/s. The circuit is basically a triode whose cathode current is fed to the anode of a pentode. There is a charging condenser between this junction and earth. A steady 10 mA current normally flows. A counter pulse turns off the triode and allows the pentode to drain current from the condenser. The next marker pulse turns off the pentode and the condenser potential remains at the value reached. This potential gives a measure of the time between pulses.

(4) A Sampling Oscilloscope is being developed. It will display waveforms with rise-time as fast as 2 ns with a sensitivity of about 25 mV/cm. It will be in the form of an add-on unit for a laboratory oscilloscope.

DISCUSSION

DE WAARD pointed out that the use of CV numbers in British articles always caused difficulties—are there no commercial equivalents? ORMAN replied that the CV408 is made by Marconi-Osram—there is an equivalent commercial number. There exists a list giving commercial equivalents of CV numbers.

WELLS suggested that the valve numbers and the manufacturer should always be given on diagrams as a measure of standardisation.

PIZER remarked that the E88CC could be used in the pulse sharpener and that the complication of the prepulse circuit was then avoided.

NUCLEAR INSTRUMENTS 2 (1958) 96–100; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION II.1

COUNTING AND STORAGE SYSTEMS

E. GATTI

Centro Informazioni Studio Esperienze, Milano — Politecnico, Milano

In nuclear physics experiments, as everyone knows, information is generally gathered through one or more particles detectors that produce electrical pulses which, with their heights and occurrence time give the desired information. This information is therefore obtained successively through one or more channels over a certain period of time in order to accumulate a sufficient number of statistical samples. The pulses have to be sorted out in classes according to pre-set requirements of pulse height, time coincidence with other pulses, etc. In fact the parallel display of information is the one needed by the physicist in order to analyze his experimental results. It is therefore obvious that as output of every class we have a final adding element (scaler, integrator etc.) where the samples belonging to a certain class are counted.

The most general type of information handling which can be done with given detectors is to "record" the experiment for instance with a multichannel tape recorder as recently done by P. E. Cavanagh and D. A. Boyce[1]) and provided this instrument is ideal we can re-run the experiment as many times as we want with no more than a loss of time. This step, permitting a possible expansion or contraction of time, merely shifts the classification of information to a later time, giving however the possibility of changes in classes if the ones chosen in a first attempt did not prove the most suitable or simply if other information is wanted from the same run.

As I shall refer mainly to immediate information processing, I mentioned the tape recording only for emphasizing that we cannot dispense, by this very useful technique, with process instrumentation. Furthermore, as the recording is not ideal and one generally loses some information, especially in time resolution, tape recording can be supplemented with some "immediate" processing: for instance the detector channels may be furnished with a recording channel fed by the output of a fast coincidence which tells us if two events which look contemporary considering the two detector tracks, are really so within a very short resolving time.

Reverting to "immediate" data processing we must have in mind the following facts: high counting rates are desired for fast gathering of accurate data, but accurate evaluation of pulse height requires periods of time comparatively long compared to those required for time evaluation (coincidence between two or more pulses). If we require coincidence it is in general convenient, to let the coincidence circuit precede the pulse height analysis in order to allow the necessary time for this operation which is thus done only on those pulses which pass through a linear gate opened by the output of the coincidence circuit. In this case, as for instance Dr. Sands pointed out at the Gatlinburg Conference[2]) one has to do with rare events, and emphasis is laid more on recording all data of these events by multichannel-multichannel operation or multihead tape recording than on speed requirements. I personally believe that the development on so-called X Y Z recording i.e. multidimensional

[1]) P. E. Cavanagh and D. A. Boyce, Rev. Sci. Instr., 27 (1956) 1028.

[2]) Nuclear Science Series Report no. 20, National Academy of Sciences, (National Research Council, Publication 467) p. 12.

pulse height analysis is one of the most interesting aspects of modern instrumentation.

If burst of pulses are to be analysed, one is forced, especially with accelerating machines, to resort to a temporary storage as made by J. Reaves[3]) on a memory cathode ray tube. This has also recently been suggested with some differences by R. V. Gåsström[4]). Another method is to store a second pulse on a memory condenser while a first is under process in the amplitude analyzer, as done by Gallagher and Johnston[5]) and Sands.

The modern trend in pulse height analysis is directed towards the use of digital computer technique to store and eventually process the information coded at the input by analog to digital conversion. This trend, initiated by D. H. Wilkinson, is largely the result of the high accuracy which can be achieved in channel width by a coding system which uses a common physical element for all channels. The lower cost per bit of stored information in comparison to individual biased channel discriminators is also important.

The disadvantage of the analog to digital conversion is a somewhat lower resolving time which however is becoming shorter with improved computer techniques or avoided with temporary storage techniques.

Having given this general picture of the problem which indicates the relative functions of coincidence, gating, and, finally, counting, I turn now to some of the ideas and instruments which recently have appeared on the scene together with some personal views. Finally I will illustrate some of my ideas and experience in the development of a 20 by 10 channel pulse analyzer with magnetic memory (which is the only storage system I am familiar with).

In the field of simple counting, the recently published scaler circuits of R. E. Bell[6]) and G. W. Hutchinson[7]) seem to be very attractive as they combine the inherent speed of an analog system of step charging of a condenser with the digital accuracy of a conventional scaler: the condenser charged by a burst of pulses is discharged to the quiescent value by pulses from a clock. The number of these charge pulses is counted by a conventional scaler. In U.S.A. a dekatron tube with ten microseconds resolving time has been announced. This makes dekatrons competitive with Philips decade tubes. Bay[8]) made a conventional binary scaler with a resolving time of 10^{-8} s and discussed the two functions of a conventional flip-flop, the one of holding the stable position and the other of switching. The function of switching is accomplished in the Bay circuit by two high current auxiliary tubes which supply in a short pulse the required currents for charging the stray capacitances.

SINGLE CHANNEL

R. D. Amado and R. Wilson[9]) developed a fast single channel analyser using two discriminators made with secondary emission tubes EFP 60 with independent thresholds. The analyser works with pulse amplitudes of a few volts and rise times down to 10^{-8} s and has a dead time of only 10^{-7} s.

LINEAR GATE

E. L. Garwin and A. S. Penfold[10]) have described a linear gate for fast pulses giving a rise time of 5×10^{-8} s and opening and closing times of the same order. They perform the switching with the beam deflection tube 6AR8.

Some preliminary experimental results I had, seem to me quite promising for the use of transistors in linear gates: it seems convenient to use the very low collector-emitter impedence of a saturated transistor in order to short-circuit a fast signal even if provided by a source of low impedence. The opening or closing time of this transistor, used as a switch, can be 10–20 ns if

[3]) See ref. [2]), p. 118.
[4]) R. V. Gåsström, Nuclear Instruments **1** (1957) 75.
[5]) See ref. [2]), p. 96.
[6]) R. E. Bell, Can. J. of Phys. **34** (1956) 563.
[7]) G. W. Hutchinson, J. of Scient. Instr. **34** (1957) 109.
[8]) Z. Bay and N. T. Grisamore, IRE trans. on electronic computers EC-5, 121 (Sept. 1956).
[9]) R. D. Amado and R. Wilson, J. of Scient. Instr. **34** (1957) 205.
[10]) E. L. Garwin and A. S. Penfolg, Rev. Sci. Instr. **28** (1957) 116.

the base is coupled by means of a parallel resistance condenser network to a low impedence gating pulse which quickly supplies or removes the storage charge.

AMPLIFIERS

The field of amplifiers this year was characterized by a widespread use of the bipolar pulse shaping technique introduced by E. Fairstein[11]) and Maeder[12]). Several commercial versions of such non-overloading non-zero shifting amplifiers are now avaliable.

It is to be noted that the bipolar pulse character is lost after the pulse strecher and therefore the circuits which follow a pulse lengthener should be d.c. coupled if minimum zero level shift has to be maintened. A linear gate should allow the passage of bipolar pulses unless the opening of the gate is a rare event. This does not decrease the value of bipolar pulses at the gate input if repetition rate before the gate is high.

A possibility which I consider useful in some instances for removing the base line shift is that of "synchronous clamping". Suppose for instance that the low impedence output E of an amplifier is coupled through a condenser C to a load R, which is normally short circuited by a switch S. Through a separate channel every pulse opens the switch S so that the pulse across load R always starts from the zero level fixed by the clamp.

Following the synchronous clamping all the couplings must obviously be d.c. couplings. Compared to bipolar shaping, the system has the drawback of not preventing pile-up of pulses in the stages preceeding the synchronous clamp. The synchronous clamp is obviously useful also as linear gate opened by signals which can be different from those of the considered amplifier.

CODING SYSTEMS

E. Maeder[13]) described a very interesting method for having a logaritmic energy scale on a gray wedge spectrometer. It is interesting to point out that the same method is useful for a logaritmic channel distribution in a multichannel analyzer which uses the Miller run down discharge of a condenser charged to the pulse amplitude; it is sufficient to let the condenser discharge in its natural exponential way instead of discharging it at constant current.

The above is but an example of an energy scale distortion which can for instance be done by modulating the discharge current of the condenser in order to suit a particular experience as advocated by Bigelow at the Gatlinburg Conference[14]).

Coding time is one of the limiting factors in analyzers using analog to digital conversion. An attempt to reduce coding time has been done in an instrument described by McMahon[15]) at Argonne in which the condenser charged to the pulse amplitude is first stepped down in 10% steps until it is within 10% of the quiescent potential. From then on it is discharged in 1% steps. The 10% steps are counted on the decade address scaler and the 1% steps on the unit scaler. This system has the drawback of some uncertainty at every tenth channel due to the difficulty of maintaining accurately a ratio of 10 between the large and small steps.

Considering pulse amplitude analyzers with magnetic core memory, it should be noticed that the usual coding process is based on the stepped discharge of a condenser. The coding process, first introduced by D. H. Wilkinson, of gating a free running oscillator is used rather rarely. I remember that in a private letter to me D. H. Wilkinson pointed out the superiority of his method compared to that using a triggered oscillator, (equivalent to the stepped discharge) in that a very narrow peak is recorded in two adjacent channels, in such proportions that the centre of gravity corresponds to the position of the narrow peak. In fig. 1 the result of calculations on the error in the centre of gravity is shown, when the analyser uses a triggered oscillator. The pulse

[11]) E. Fairstein, Rev. Sci. Instr. 27 (1956) 475; and ref. [2]), p. 21.

[13]) See ref. [2]), p. 27.

[13]) E. Maeder, see ref. [2]), p. 43.

[14]) J. H. Bigelow, see ref. [2]), p. 143.

[15]) J. McMahon, see ref. [2]), p. 154.

distribution in the peak is assumed to be triangular. The parameter a is the base of the triangle in units of the channel width. The abscissa is the true position of the peak, also in units of the channel width.

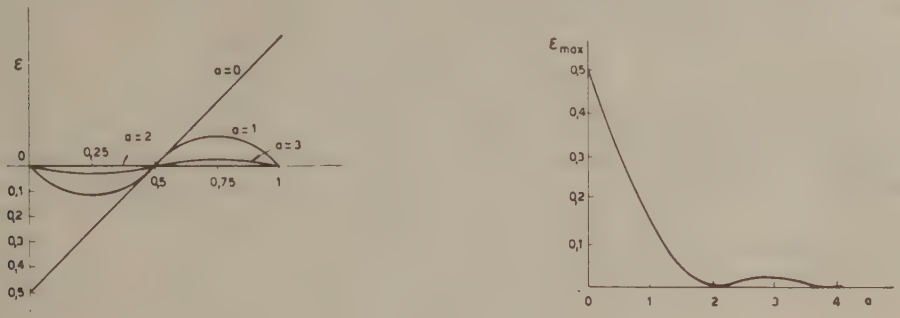


Fig. 1

It is obvious that the errors are negligible when the peak is larger than 3 channels, and that (apart from α-spectroscopy where the peaks can be very narrow, even one per cent) the two coding systems are equivalent. The gated oscillator method is however better when few channels are used. Furthermore, this method is by far the most convenient for ascertaining instrument accuracy by determining the position of a narrow artificial calibration peak from the counts in two adjacent channels. A drift in its position, even of a small fraction of a channel, is easily detected.

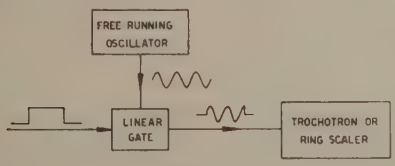


Fig. 2

It is useful to draw attention to a possible error which can occur when a gated oscillator is used to drive ring scalers or trochotrons. This error is illustrated by fig. 2.

The fraction of the last cycle, which is mutilated by the gate, can cause a transition of the address circuit or not, depending on the sensitivity of the ring scaler or trochotron at its particular position: different sensitivity for different positions will cause differences in channel width. Errors are reduced if, istead of the sinusoidal wave, narrow pulses are used as P. R. Shifflet[16]) did, but it is even better to introduce a trigger circuit between the gate and the counter. This trigger circuit should have very short recovering time so that its sensitivity is not affected by the number of preceeding pulses.

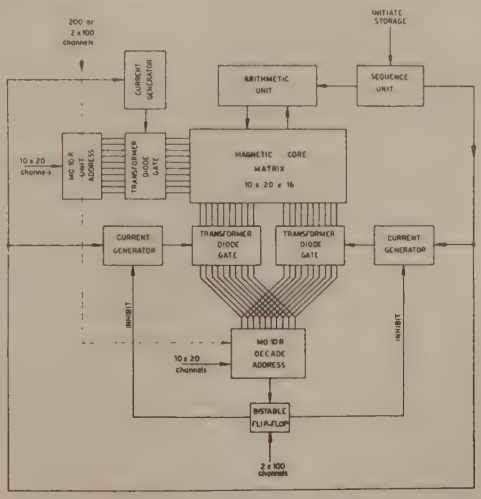


Fig. 3

STORAGE

Among so many specialists I do not find myself qualified to speak in a general way about storage, as my experience is limited to magnetic core memory and even in this field my

[16]) P. R. Shifflet, see ref. [2]), p. 95.

experience is quite recent. So, I will briefly refer to the development under way in my laboratory.

We are developing a magnetic core memory with 16 10 × 20 matrices having in mind to build a universal memory unit not necessarily connected with a definite coding unit. All circuits are transistorised except the trochotron address circuits.

Details about the core driving circuits for rows and columns will be given in the next section. We designed the memory for operation either as a 200 channel analyzer, as a 2 × 100 channel analyzer with two independent inputs for two simultaneous experiments and finally for 10 × 20 channel in X Y Z operation. With a reduction of the counting capacity 20 × 20 channels can also be provided.

A simplified block diagram is given in fig. 3.

We have 16 μs spent on reading, adding and writing processes plus a counting time of half a microsecond per channel. We use cores with 1 μs switching time and OC44 transistors except in the address current generators where 2N269 are also used.

DISCUSSION

HUTCHINSON: The advantage in calibrating a machine with floating channel boundaries in place of stationary channel boundaries is rather illusory. In fact, if you have fixed channel boundaries, the method of calibrating a particular channel boundary is to slide a pulse of standard amplitude up the channel until it counts equally in two channels. That is exactly equivalent to the weighting process that you get with floating channel boundaries and I think it is at least as accurate.

GATTI: I agree, as far as accuracy is involved. However, such measurements require a very accurate setting of the calibration pulse or of the bias of the pulse height analyzer.

Of course every fixed boundary instrument could be modified to a floating channel boundary instrument merely by superimposing a triangular waveform with an amplitude of one channel width on the incoming pulses.

HUTCHINSON: I am a little suspicious about the argument of the floating boundaries, because obviously if you float the boundaries, you are losing some fundamental information.

GATTI: I think you loose some fundamental information if you do not know where a very narrow peak is.

HUTCHINSON: This is not a very realistic problem in physics.

GATTI: I think on the contrary that it is a realistic problem for instance in X Y Z operation, where one has only a few channels, say 10, for each variable: in these instances a peak can well lie entirely in one channel and its energy cannot be accurately determined.

DARDEL: The difference between the floating and fixed boundary is that in the second case you have a rectangular and in the first a triangular window curve which has somewhat lower power of resolution. The floating boundary has the advantage that you use a continuously running oscillator which is less likely to introduce systematic errors.

KANDIAH: During the 2 × 100 channel use of the instrument you have shown on the block-diagram, would the common arithmetic circuits used there prevent you from storing in one half of the equipment while the other half is processing some information? If for example, the first 100 channels received some information, the address register selected an address and the arithmetic circuit was was processing the information at that address, then the store is not accessible to the other half.

GATTI: When working with two separate inputs for operation with 2 × 100 channels, one has a dead time due to the other experiment unless a temporary store not yet built is provided. I could make an analog temporary storage of the incoming pulse of the other channel.

KANDIAH: Except that you will still from the viewpoint of any one user, look into a machine which is dead for a period of time which is not known to him.

GATTI: Yes: but in many instances this is not too much trouble if the total dead time is measured. With random pulses it is not necessary to distinguish between dead times due to the one or to the other experiment.

NUCLEAR INSTRUMENTS 2 (1958) 101–108; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION II.2

SERIAL COMPUTER TECHNIQUES

G. G. SCARROTT

Ferranti, Manchester

I am a little diffident about talking to nuclear physicists because I have not had anything to do with nuclear physics instruments for some years. So I am just going to talk about some computer techniques in the hope that they may be of some interest.

Although you probably know that computers require regular maintenance to give good service, in fact they are like this simply because they represent an amount of electronics which is simply enormous compared with even the instruments that you physicists use, and it is really rather a miracle that they work at all. To achieve the present reliability, the quality of the individual computing elements has to be extremely good; and therefore the details of the circuit have to be designed in an extremely conservative way. A performance of the individual circuit elements which would be tolerable in most nuclear instruments would be quite intolerable in a computer because the complete computer would only work for 5 minutes at a time. This problem was very obvious when the first computers were made because they were designed by people who were at once mathematicians, logical designers and circuit engineers. In the course of doing all these three quite difficult jobs at once, the detailed circuit designs sometimes became such that it was quite impossible to analyse their behaviour under all possible conditions of component drift and stored patterns, and as a result, sooner or later, some peculiar combination of circumstances would happen which just made them not work.

Of course, the solution to this problem has now become to be recognized everywhere that one must not do the job of circuit design—and logical or computer design—at the same time by the same people. And so these days people all try to design computers by designing logical elements which are what you call "gates" and make these gates have very well prescribed properties. They are such simple devices that the circuit designers can test them under all possible conditions and their properties are sufficiently flexible to permit the logical designer to carry on with the design of the computer without knowing how the gates work, without necessarily even caring how the gates work.

The subject of my talk is serial computing techniques and at this point I had better underline the difference between serial and parallel computing techniques. You probably know this anyway but it is worth stating that if a number is represented by a pulse pattern coming along one wire, this is the serial method of representing it, whereas if it is represented by single pulses coming simultanously along many wires, this is a parallel representation. One might think at first sight that the logical elements required for handling the information would be exactly the same whether the information is in serial or parallel form and that therefore there is no point in bringing in the question. In fact, I do not think this is true because any logical elements has 2 characteristic times which are associated with it:

(1). One is the propagation time, which is: how long after the inputs have been fed into it does the output become available.

(2). The other one is the resolution time, which is: how soon after an input signal has been fed into it can one dare to put another input signal to it.

One might call these times T_p and T_r and the ratio of these two times T_p/T_r is a number which

indicates something of the properties of the logical element.

Now, in serial computers, there is no hurry to produce carries because when two numbers are added together the carry comes out straight away, although it is not strictly required until the next digit comes along. Therefore, there is no hurry to get the answer and the logical designer is perfectly happy to have the ratio T_p/T_r about 1 in elements for a serial computer.

In a parallel computer, however, unless the propagation of carries between digits is made very much faster than the resolution time between numbers, the parallel mode does not yield a useful speed advantage over the serial mode. So that for a paralled computer element $T_p/T_r \ll 1$. There is therefore a very real difference between logical elements for parallel computers and logical elements for serial computers.

It happens that all the work with which I have been associated in the last 3 or 4 years has been concerned with serial computers, and hence the logical elements that I am going to talk about are not well suited for making parallel computers.

Another point which is worth stating, especially as we have been talking so glibly about nanoseconds, is that we have been aiming at a good engineering compromise all the time; we were never trying to go as fast as we possibly could regardless of expense. We were always trying to make a device which would treat most information most cheaply. I shall therefore be talking about time scales which are rather slow compared with those I have been hearing about in this conference.

So having specified roughly what a serial logical computing device is, we can now fill in a few more details. Most computers are built with logical elements made with diodes and which are fundamentally exactly the same as the gates you use in nuclear physics.

The elements are plugged in or connected up by some comparatively cheap systematic method and therefore several elements are provided on each plug-in unit, since otherwise the plugs and sockets may contribute more nuisance value than the circuit components. The design problem therfore, is to allocate the various kinds of gate on the computing element in such a way that there is a minimum of unused gates in the final computer. One, therefore, has to think very seriously about how to make logical elements which are as flexible as possible in their use.

We have done some work on such an element, and in order to explain this work it is necessary to make some assumptions which may sound arbitrary. One of these is that we have used a current to represent a digital variable rather than the more conventional voltage. This assumption is not strictly necessary to exploit the principle I am going to explain, but it does have certain advantages of convenience.

The use of a current as the prime variable implies that logical elements have a low input impedance and high output impedance and if a variable is to be fed to several elements they must all be connected in series. It is quite easy to make a summation transformer (fig. 1) which

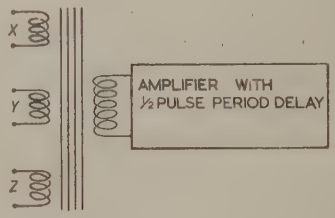


Fig. 1

can linearly add currents, i.e. if we put in X, Y, Z, where X, Y, Z are all Boolean variables, then the use of the transformer has conferred great flexibility because the current that comes out can be all sorts of functions of X, Y, Z, depending on which way the windings are connected. A great variety of logical functions can be made available simply by connecting the output of the summation current transformer to a device which provides power gain with a threshold. The final output is then a current source of value one, if the weighted linear sum of X, Y and Z is greater than the threshold, but is zero if the sum is less than the threshold. The

output current is, of course, in a form suitable for direct connection to the inputs of other logical elements.

There is one other matter which should be mentioned. It is pretty obviously fundamental that no logical device can give an output at the same time as it receives its input. It must have an inherent delay and we have decided to increase the delay to a standard value of one half a pulse period, thus making the ratio T_p/T_r for our logical elements have a ratio of $\frac{1}{2}$.

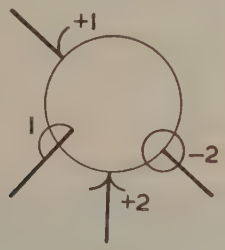


Fig. 2

With this value of delay it is possible to restandardise the pulse out of every logical element. The delay must of course be taken into account when drawing the logical diagram, but provided this is done there can never be difficulties due to inadequate restandardisation.

Having defined the logical properties and the time and amplitude restandardisation properties

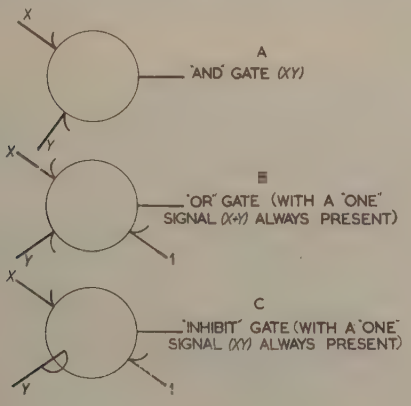


Fig. 3

of the logical elements, we can now show how useful these devices are. Fig. 2, shows the symbolism employed to permit the drawing of logical diagrams.

Inputs are drawn with an arrow, or half an arrow, to denote input windings connected to yield a weighting factor of $+2$ or $+1$. Similarly an input with a circle, or half circle, represents a weight of -2 or -1.

The diagrams 3A, 3B and 3C, show three ways of using the elements to make an "and" gate, "or" gate and "inhibit" gate respectively. These all use the same device with different connections in the wiring at the back of the sockets so that a logical element does not know what task it is to perform until it has been plugged in. As an example of the application of these devices fig. 4 shows a logical net which is able to add one

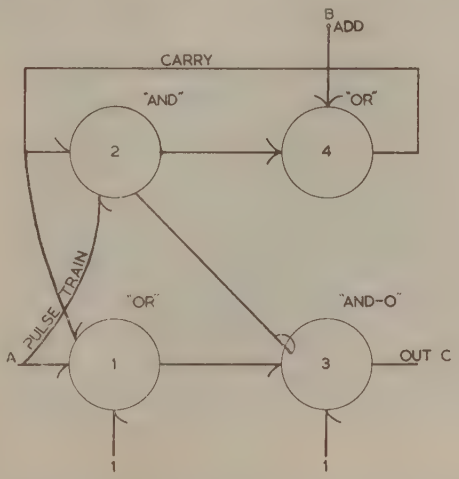


Fig. 4

to a serial binary number. A pulse train representing a number is fed in at "A", and a single pulse at "B". The sum comes out of C and the net, collectively known as a half adder, uses 4 elements.

The engineering details of the threshold amplifier determine the position of the threshold, and I will briefly describe one form of amplifier which conveniently makes the threshold have a value of $1\frac{1}{2}$.

If a square-loop ferrite core is equipped with two windings and a current very large compared with the critical field is put through one of the

windings; the secondary winding, if connected to a low impedance, will pass a current which is closely proportional to the primary current, as long as the core is in its unsaturated state. In fig. 5 there are two such devices; I is the large current we shall call the shift current. The resultant secondary current has a well defined value as explained which is able to switch another core when passed through its input winding. The diode between the cores is connected with such polarity that it does not conduct when a core is being switched by its input winding, but does conduct when it is being switched back by the shift current.

If the core is equipped with 2 or 3 windings on it instead of 1 then it can easily be arranged that one current by itself is not sufficient to

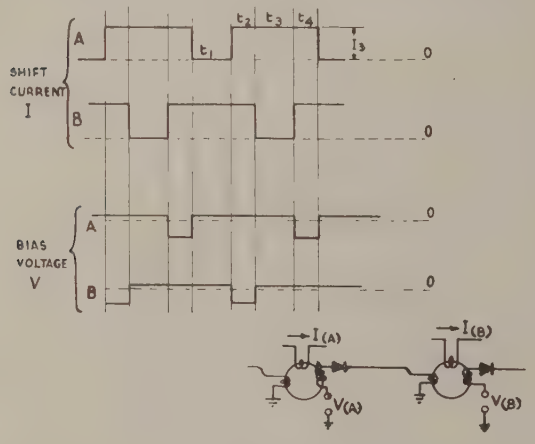


Fig. 5

switch the core but two is, which therefore means we have a threshold of $1\frac{1}{2}$.

The next design feature which we have to discuss is the property which must be built into logical elements to make the machine a binary machine.

This property which defines a binary logical element is the fact that a curve plotted for its input against output must look like fig. 6A. It is a binary element because the curve has 2 kinks in it.

It is no use having the transfer curve like fig. 6B for example, because in that case the output after having been propagated through many stages might be any value other than the desired quantised values of 0 and 1. There is a

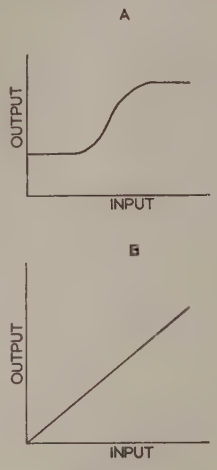


Fig. 6

very easy way of tailoring the transfer curve which is shown in fig. 5. It is very obvious that the time when the current comes out must be different from the time when the current carrying information went in. So therefore, the elements are inevitably two-phase devices. All devices are driven either by an A phase or B phase shift signal. Then A outputs can be fed to B inputs and vice-versa, but A outputs can not be fed to A inputs.

We have to define how these A and B phase shift currents are going to be related to one another.

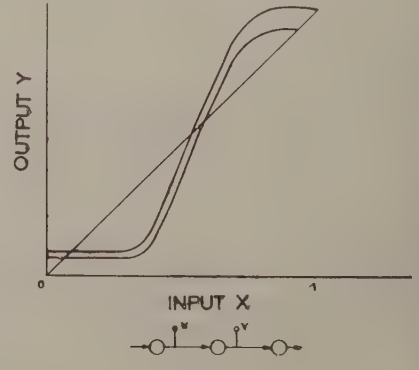


Fig. 7

Each phase current is arranged to be turned on for more than half the time, so that there is a

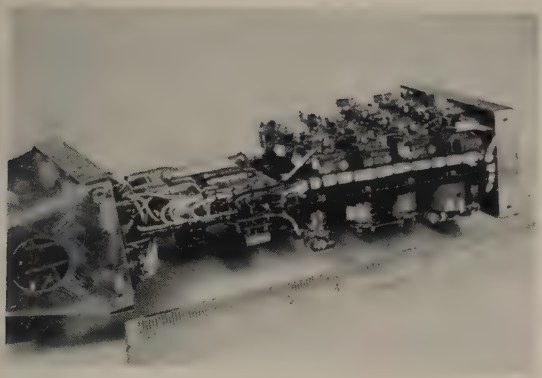


Fig. 17

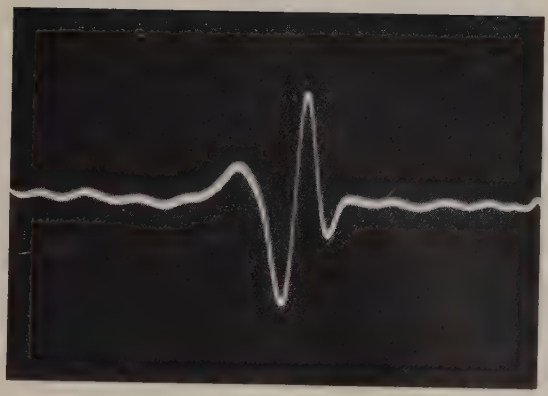


Fig. 18

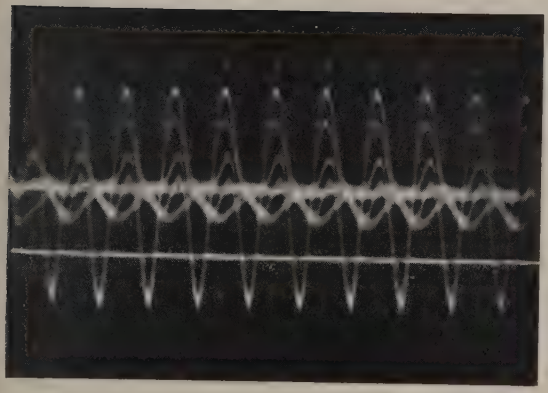


Fig. 19

Fig. 20

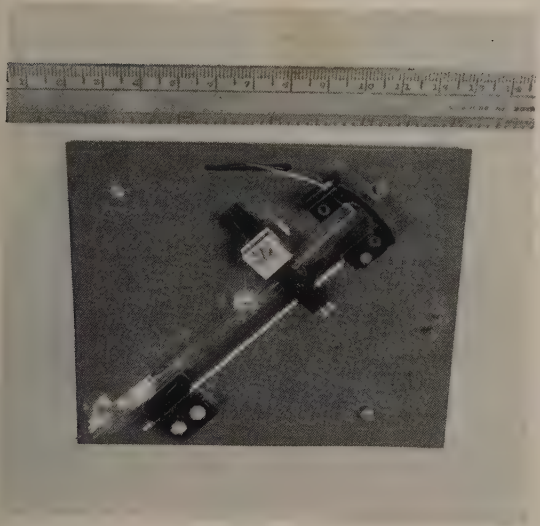


Fig. 21

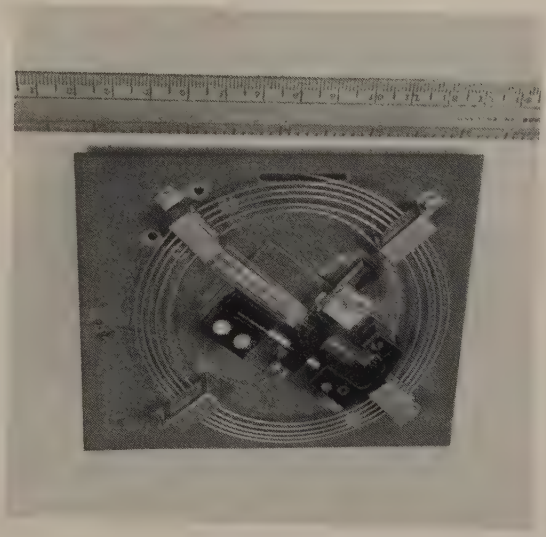


Fig. 22

time when current is on in both phases. Thus, if a small amount of flux is stored in the first core, then the resultant output current cannot last sufficiently long to feed an input to the next core since the second core can accept information only when it has zero shift current. Fig. 7 depicts an experimental transfer curve plotted using flux stored in two consecutive cores as the abscissa and ordinate and shows that the use of overlapped shift pulses does in fact yield the desired transfer curve with two kinks.

The fig. 8 shows the mechanism whereby the whole device is a power amplifier. This is a diagram which relates the gross mmf from the shift and input windings to the flux in the core.

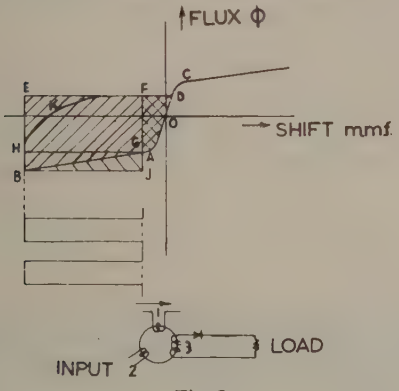


Fig. 8

The product of these two quantities is energy. When information is put in, the state changes from G to D and the generated EMF is of a polarity which leaves the diode cut off so the work done is the area GODF and is very small. When the shift winding is excited the state of the core as seen from the point of view of the shift winding goes from D back to G via the path DEBG, since the diode conducts and in effect makes the core appear more lossy due to the presence of the load resistor. Thus a large energy DEBG is dissipated in the load conditional upon the supply of a small input energy GODF.

Fig. 9 shows the transfer characteristics of an element used as an "and" gate. There is a family of curves since there are two independent variables and the form of the family shows that the transfer characteristic has the required state for both the possible inputs.

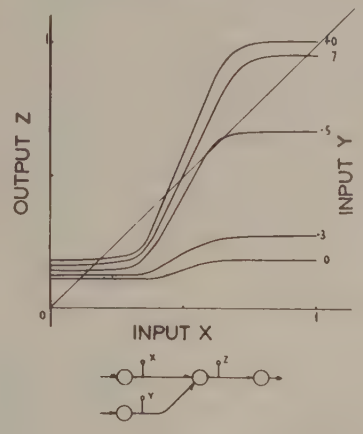


Fig. 9

Fig. 10 shows a number of the logical functions which are permissible in a single element.

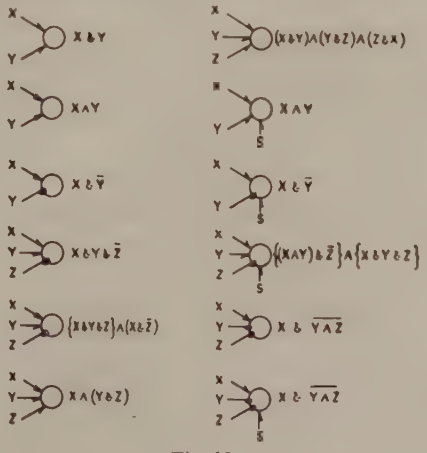


Fig. 10

Fig. 11 shows a number of alternative forms of serial binary half adder which are derived by reversing the sign of the various internal connections, leaving the external properties unaffected.

Delay Line Storage

Another typical serial technique is the use of delay line storage.

There are two kinds of delay line storage. Acoustic delay lines and electro-magnetic delay lines.

An acoustic delay line usually causes a very large attenuation and therefore requires a large amplifier. An electro-magnetic delay line, on the other hand, causes very little attenuation, but cannot easily be made to store much information. There is, therefore, a clear division of application in that large scale storage is achieved using acoustic delay lines, whereas, small scale storage of up to about 10 to 20 bits may be done by the use of electro-magnetic delay lines.

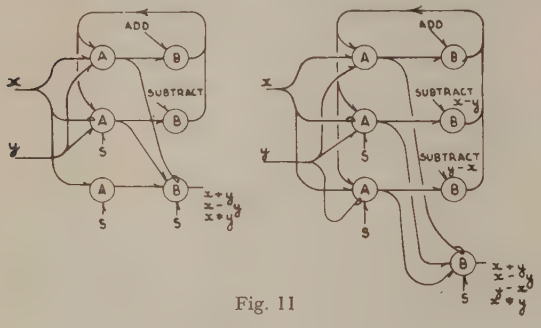


Fig. 11

You are probably only too well aware of the fact that ladder type delay lines never work as well as one expects and, in fact, there is a very good reason for this difficulty.

If a ladder type delay network is desinged to have a linear phase delay, i.e. a time delay independent of frequency, then its amplitude frequency response is inevitably of Gaussian

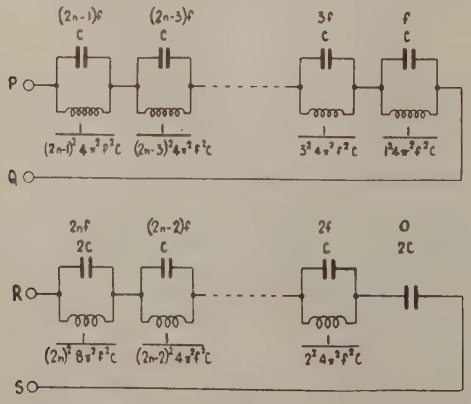


Fig. 12

form. This means that at best the number of digits stored in such a line is proportional to the square root of the number of components.

Therefore, the problem arises how can we make an electro-magnetic delay network other than by using a ladder network. Now a network consisting (fig. 12) of a series of parallel tuned circuits, tuned to frequencies in the ratio 1, 3, 5 etc. behaves like a short circuited delay line. A similar network with the circuits tuned in ratio 2, 4, 6, etc. behaves like an open circuit delay line.

Now it is well known that one can connect, two of type A and two of Type B to make a lattice type delay network which uses many components because two of each type of network are necessary. However, a lattice network provides a storage capacity proportional to the number of components. By the use of a valve to complete the bridge it is possible to make a network with the properties of a lattice delay network but with only half the components.

The fig. 13 shows how it can work. The two networks generate equal and opposite signals

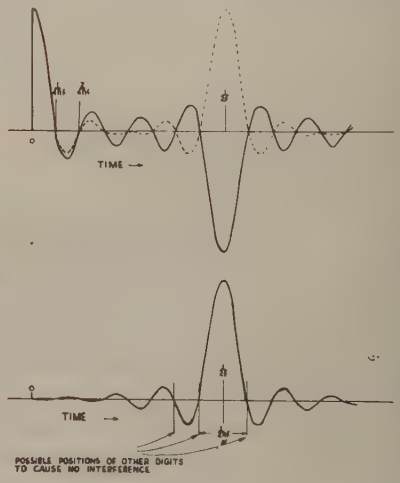


Fig. 13

when an input is fed into the valve and so the bridge is balanced and no output results. When the reflected signals appear after a delay one is reversed and the other is not reversed, and hence, the reflected signals do not balance and an output signal results.

Fig. 13a shows the signals from the individual networks and 13b shows the output of the bridge.

Fig. 14 shows one practical form of the valve bridge circuit used as a lattice delay network.

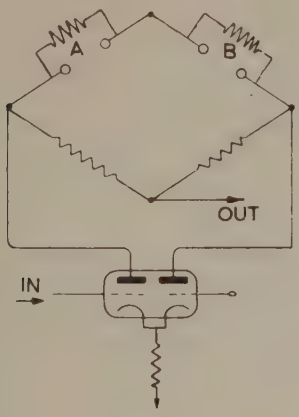


Fig. 14

Fig. 15 is a photograph of a cathode-ray tube screen showing many of the possible digit patterns in a 28 bit-line superimposed on one another by using a time exposure on a rapidly changing digit pattern. In an ideal delay network the ripple in the signal which preceeds and

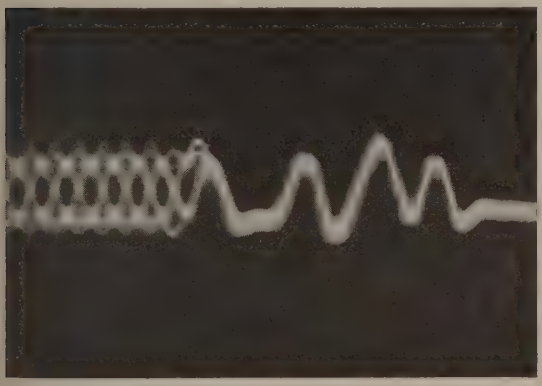


Fig. 15

succeeds the pulse has a value zero at all other possible pulse positions, and hence, at every pulse position the output voltage can have only two values. The photograph of the output of such a line will therefore show two bright spots at every digit time and fig. 15 shows that this is indeed so, even when the line is storing 28 bits.

Turning now to acoustic delay lines for large scale storage. I will refer briefly to some work on wire type delay lines. After a few experiments we came to the conclusion that conventional wire delay lines using longitudinal waves suffer from dispersion which seriously reduces their storage capacity. The chief cause of this is an effect due to the curvature of the wire which is inevitably present since a practical line, containing several meters of wire, must be fitted into a reasonable space. This effect causes dispersion of the low frequency components.

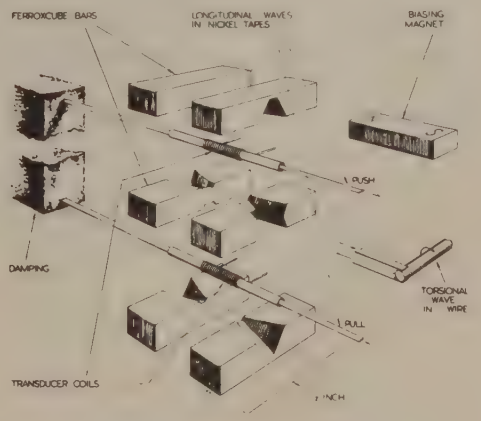


Fig. 16

Another dispersion mechanism affects the high frequency components due to the coupling, via Poisson's ratio, between longitudinal and transverse modes of vibration. So that in fact there are two reasons why longitudinal mode delay lines do not work very well.

Neither of these reasons, it turns out, apply to torsional mode lines. And so the problem became: how can we make torsional waves?

Fig. 16 shows how to do this. Two pieces of nickel tape are welded tangentially to the wire. Each is fitted with a coil and a polarising magnet is arranged to magnetise both tapes in the vicinity of the coils. The coils are connected in series with polarity such that when current is passed one piece of nickel tape has its magnetisation increased and in the other the magnetisation is decreased. Hence, a positive

longitudinal stress wave is generated in one nickel strip and a negative one in the other. These arrive simultanously at the junction with the wire and produce a torisonal wave in the wire. The same device at the receiving end permits the detection of the torsional waves.

Fig. 17 shows 2 delay lines made according to this technique. Fig. 18 shows what a pulse looks like after it has been through 2 000 digits delay (1 digit = 2 μsec). Fig. 19 shows what many superimposed patterns look like after going through 2 000 digits delay. Fig. 20 shows a printed circuit amplifier for the delay line. Fig. 21 shows a delay line which holds some 300 digits. It is very small, 5″ square, and can be attached directly to the amplifier of fig. 20. Fig. 22 shows the other side of the same delay line. We put one transducer on one side and another one on the other. By this device, we can arrange that the line can be cut to any length without any changes in the design.

By the use of these logical and storage elements it is possible to reduce the design of most of the internal parts of a computer to an exercise in logic. The resulting freedom from purely engineering worries can lead to greater complexity in the operation which, in turn, permits the design of computers which are easy to use from the programming point of view.

NUCLEAR INSTRUMENTS 2 (1958) 109–113; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION II.3

RECENT DEVELOPMENTS IN COUNTING EQUIPMENT

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The main developments at Harwell have been connected with the utilisation of new electronic devices in order to improve the reliability of electronic equipment. We have carried out periodic checks on reliability using the records which are kept in the Establishment of all failures of equipment designed by the Electronics Division. It has been clear for a long time that the reliability of electronic equipment is far from adequate. The general pattern of the causes of unreliability has remained unchanged over many years.

Thermionic valves have been one of the major factors. Their life is not adequate and they dissipate power and thus decrease the reliability of other components. We have therefore attempted to design a new range of equipment using transistors and magnetic cores as much as possible. We expect a reasonable improvement in reliability and in some instances improved features in the equipment. I shall now describe a few of these developments.

1. Slow Scalers

In view of the need to record automatically the number stored in a scaler we have not considered mechanical registers.

The dekatron is one of the simplest and cheapest of scaling devices. They are inherently very reliable but have occasionally been maligned owing to the limitations of the driving systems. Some of the driving systems do not make allowance for the following:

(a) There must be adequate overlap between pulses to the A and B guides—cross-over at below—30V relative to the main cathodes.

(b) A change of maintaining potential takes place on cathodes adjacent to that on which the glow has been resting continuously.

A common pulse driving system using one cold cathode tube per dekatron which allows for these effects has shown good life.

A transistor drive circuit meeting these requirements has been developed. It uses 1 transistor, 5 diodes, 5 resistors, 2 capacitors and a transformer. If the resolving time of the scaling stage could have a tolerance of $\pm$ 20% the circuit would work without fail with components—resistors, inductors, capacitors—varying by 20%.

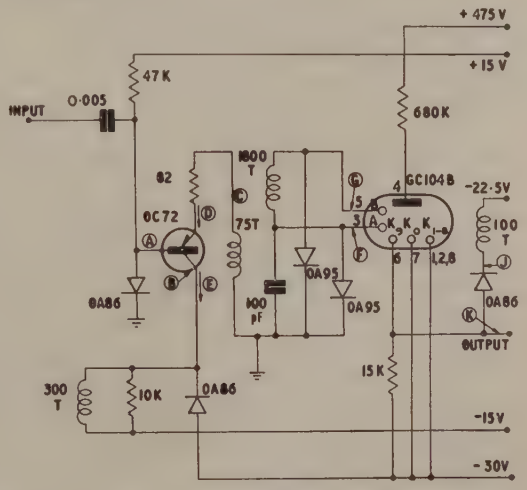

Fig. 1. Transistor driven dekatron (intermediate stage).

Fig. 1 shows the circuit diagram. The transformer with 4 windings is on a ferroxcube core with a nominal tolerance of 20% on permeability. The transistor is used in a blocking oscillator circuit with a resistor in the emitter lead and the inductance determining the pulse

duration—independent of supply voltages. The pulse amplitude is proportional to the supply voltage.

The transistor collector bottoms for a time kL/R. It then overswings to twice the voltage and is caught on the diode for the same duration. It then recovers critically damped to the supply voltages.

The 100 pF capacitor from the A guide to earth ensures that there is an overlap between the A and B pulses.

Using low frequency transistors a resolving time of 300 μs is reliably obtained.

One major advantage of this circuit is the small delay to the carry pulse—which is of the order of 3 to 4 μs per stage.

Fig. 2 shows the waveforms in the above circuit.

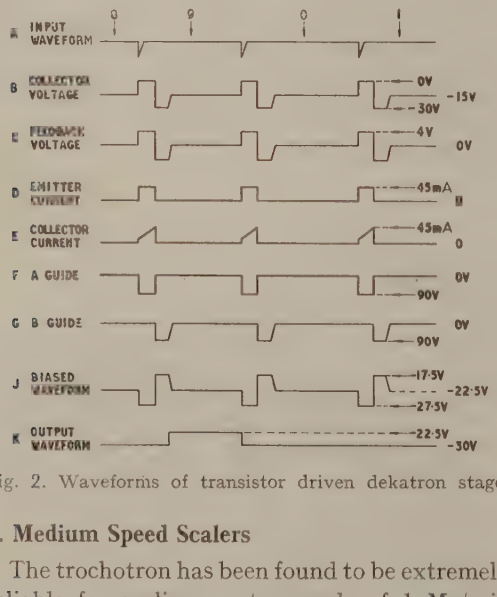

Fig. 2. Waveforms of transistor driven dekatron stage.

2. Medium Speed Scalers

The trochotron has been found to be extremely reliable for scaling up to speeds of 1 Mc/s in completely aperiodic circuits. It requires 2 valves to drive it reliably at this speed. What is more important, the valves are connected in a bistable circuit and therefore require the usual collection of resistors with limits on their values. Ignoring the known failure rate of such bistable circuits the trochotron is very reliable.

In the case of the trochotron again the only known cases of unreliability are due to incorrect use of the trochotron. It is essential to have:

(1) No overlap between pulses.

(2) Low enough d.c. resistance in the switching grid circuits particularly at high spade supply voltages.

(3) Optimum spade loads for a given spade voltage.

It should be remembered that under favourable circumstances the time for a trochotron to lock down is less than 0.15 μs.

A binary circuit which ensures this uses an inductance as the memory element and requires only one cross-coupling from anode to grid with critical components. The advantage of this binary is that it has no overlap and is relatively easy to switch.

A more elaborate version of this binary using pentodes and a delay line memory is capable of a resolving time of 50 μs.

The trochotrons with built-in resistors will resolve 0.1 μs when driven with two pentodes following the above binary. But there is some uncertainty about the future of this fast scaler since it will be possible to obtain this speed with transistors expected to go into production shortly.

3. Ferrite Core-Transistor Scaler (Dr. E. FRANKLIN)

We have been considering scalers using square loop ferrite cores as the memory elements driven by transistors because of the high reliability to be expected of such a system. The system using a ring of 10 cores requires critical pulse amplitude and duration. The product of pulse amplitude and duration should be constant within $\pm$ 15%. In a preferred system one would like a limit only on one side—that is that a quantity such a pulse amplitude or duration should be merely greater than or less than some value.

The system using two sets of 5 cores is more satisfactory from this point of view. In particular the system using three windings per core is particularly attractive. In this system there is a two way drive circuit. One drive circuit switches one of the cores 0, 2, 4, 6, 8 and the other switches one of the other set of cores.

Only one of the cores is in the "1" state. When a pulse is applied to switch this core to the "0" a "1" is set up in the next core.

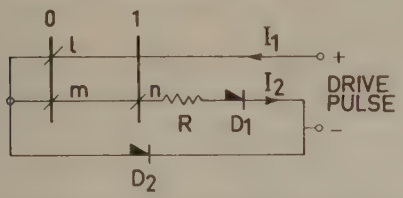

Fig. 3. Single stage of the improved two-phase circuit.

Fig. 3 shows the connections of two adjacent cores. Each core has windings *l*, *m* and *n*. The current in the winding *n* of a core sets it to "1". The voltage developed across *m* when it is set to "0" drives the next core (*n* winding). The difference between the currents in 1 and *m* sets the core to "0". The diode D_2 is cut off during switching by the voltage developed across *m*.

the path of the current in the lead past the *l* windings.

The transistor binary uses similar cores as memory elements. Both transistors are normally cut off. When an input pulse is applied one of the transistors, depending on the state of the cores A and B, is switched on. This applies switching current to one or the other sets of *l* windings. The transistor circuit switches off only after the appropriate cores in the scale-of-ten have fully switched.

The order of completion of switching is

(a) The core driven on the *n* winding.

(b) The core driven on the *l*–*m* winding.

(c) The A (or B) core.

There is a minimum limit to the supply voltage below which the system will not switch. This is around 10 volts. The lowest permissible supply is set at 12 volts. The upper limit is set by the transistor. Increasing the

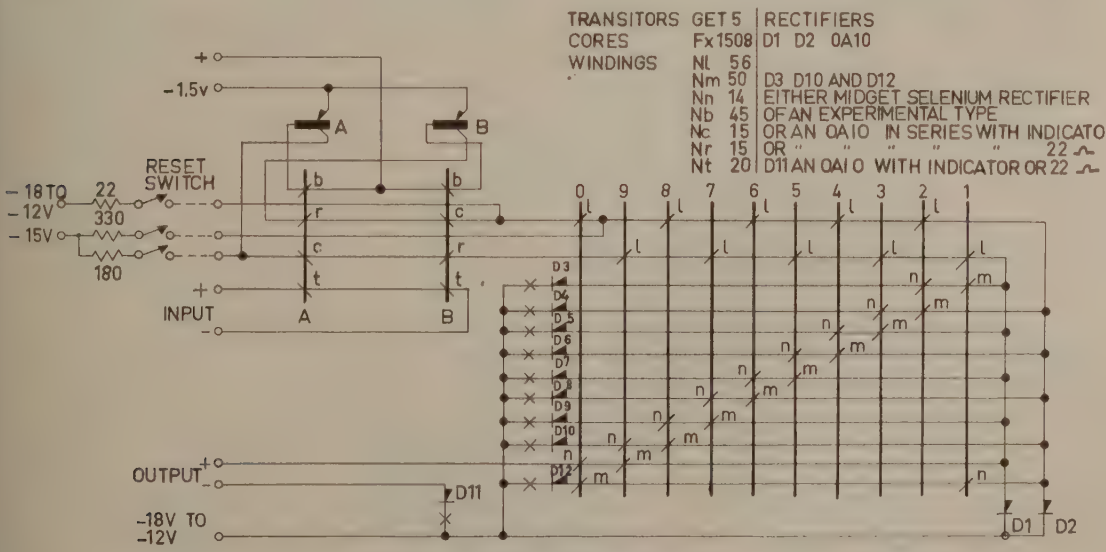

Fig. 4. Improved two-phase circuit.

Fig. 4 shows a complete scale-of-ten using this system. Note that the winding *l*, *m* and *n* are 56, 50 and 14 turns respectively. This ensures that a "1" is set up on a succeeding core before the driving core is switched to a "0". The voltage induced into the *m* winding determines voltage merely reduces the switching time of all the cores. The circuits have operated satisfactorily at over 20 V with transistors that could withstand this voltage.

The transistors used have a nominal α cut-off of 1.5 Mc/s and a power rating of 125 mW. A

resolving time of 3 μs is obtained with these transistors. The maximum mean rate, limited

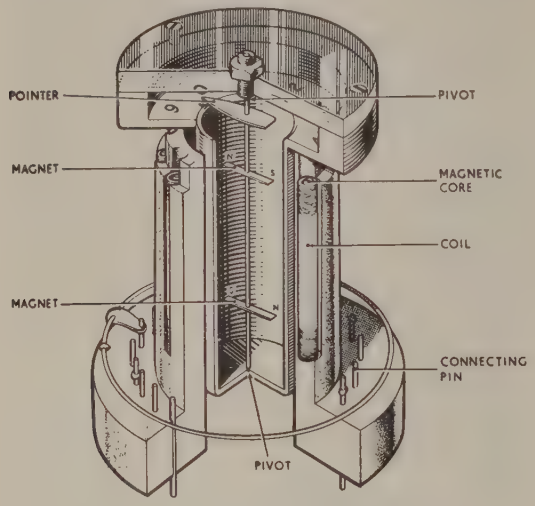

Fig. 5. Arrangement of magnetic indicator.

purely by transistor ratings, is about 250 kc/s.

With transistors with f_a of about 5 Mc/s a resolving time of 1.0 μs has been achieved but because of the low power rating of these transistors the maximum frequency is less than 100 kc/s.

An indicator using square loop strips magnetised by the core currents has been developed. A cut-away view of this is shown in fig. 5. It uses a pair of astatic magnets inside an arrangement of 10 magnetic strips which get magnetised by the switching currents in the scale-of-ten. The response time of this indicator is about 1 s.

4. Analogue-to-Digital Converter

Fig. 6 shows the essentials of the new analogue-to-digital converter used in the pulse analyser being built at Harwell. It is worth noting that this converter consumes about 150 W, the address circuits of the analyser consume about 80 W and all the remaining circuits consume about 40 W. The efficiency of the power supply is around 60% overall.

The essentials of this analogue-to-digital

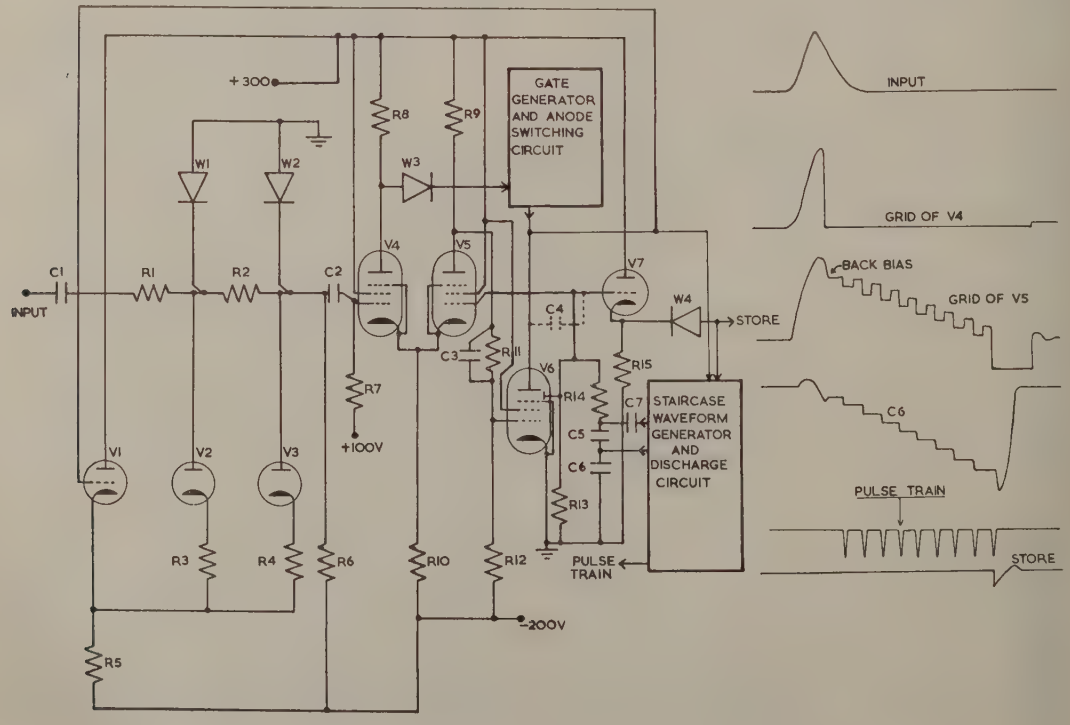

Fig. 6. Block schematic and waveforms of pulse analyser encoder.

converter are that the circuit that is used for stretching the pulse is also used at a later time to compare the stretched pulse amplitude with the output of a staircase waveform generator. The circuit acts as a negative feedback system during charging and as a positive feedback system during the comparison with the staircase waveform. This is accomplished by reducing the anode voltage of the secondary emission pentode to below the dynode voltage during the application of the staircase waveform.

A channel width of 100 mV with a stability of 1% can be readily achieved with this system. d.c. drifts are relatively unimportant except in the circuit that determines the magnitudes of the step in the staircase. Channel boundary stability of the first few channels is a few millivolts over some hours. The application of back bias is relatively simple. This is done by applying the negative step at the anode of the secondary emission valve, suitably attenuated, through a fixed capacitor to the storage capacitor.

NUCLEAR INSTRUMENTS 2 (1958) 114–116; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION II.4

SOME REMARKS ABOUT PULSE-HEIGHT ANALYSER DESIGN

G. W. HUTCHINSON

University of Birmingham

I would like to make a few remarks about what I consider should be the trends of a pulse-height analyser design. I think it is important to discuss this because pulse-height analysers are going to be used in many laboratories in the world and are going to be made in large numbers; and, if at this time, the wrong decisions are made about the type of analyser which should be widely made, then it is going to cost quite a lot of money.

Now I am talking about multichannel analysers of the order of 100 channels and of not very high speeds—and I should like to come back to the point of speed a little later on. If we consider only serious instruments where the accuracy is required to be considerable, it is clear, I think, and everybody would agree, that we have to use a digital storage system. This means the techniques that we have to employ are exactly the techniques of digital computers and the circuits which will be used will be the sort of circuits that we usually put into computers. Of these circuits the main choice is to be made between the parallel and the serial systems.

In the original analyser that Scarrott and I made in Cambridge, we used a serial system and I think we had good reason to do so. But at the time when we made it, electronics was not nearly as advanced as it is now. We did not know much about either system, as a matter of fact, and it may be that our reasons were not the right ones. I would therefore like to open the general discussion of which of these two sorts of system one should choose. And it seems to me that the basis on which one should discuss this is the following.

In digital computers themselves a great deal of research has been put into the memory systems and in fact in various parts of the world both types of memory are now used. Whether the balance is in favor of parallel or serial systems, I would not like to say. I am not in that field. But the fact remains that both sorts of system are used. It is therefore to be presumed that both sorts of system (the one exemplified by the accoustical delay line, and the other by the matrix ring store) have a similar balance of reliability, general convenience and expense.

One should therefore look, in designing a pulse height analyser, for an auxiliary criterion, additional to those relevant in making digital computers, which may seriously upset this balance.

I think this extra factor is that, in a pulse height analyser, you are only adding one thing at a time to an existing store, and if you are using any form of height-to-time conversion, the position of the new information in the store is determined by the act of measuring the pulse height. Thus access is only required to one part of the store at a time, and whether a serial or parallel system is used, the appropriate part of the store is always immediately accessible at the time when the height of a pulse has just been measured. The easier access normally associated with parallel systems is therefore not useful in this case and the speed of the analyser is essentially governed by the comparison process.

This is not a new argument. It has also been made by Mr. Scarrott. It means that the usual justification for a parallel system fails in the particular case of a pulse-height analyser using height-to-time conversion.

One can ask then if there is any other argument in favor of the magnetic cores; because clearly, if there is not, they are more expensive and elaborate, and certainly for a home-made instrument (and I think also for a commercially made instrument) the serial memory is a much easier one to construct.

One answer that has sometimes, I think, been given is that the magnetic cores are intrinsically more reliable.

I think this is now a misconception. I am afraid that Scarrott and I did one disservice to nuclear instrumentation when we built our first kicksorter; but this was no fault of ours. What happened was that we were very unfortunate in having an extremely good bit of nickel wire which was lying around the Cavendish; and the fact is that when we first made nickel delay lines with this wire, we simply put two little coils on the end of them and they worked immediately. We had no trouble and we went straight forward. It was only a few months later that we found troubles. Many other people have found them since, and the fact that other bits of nickel wire do not work with such simple treatment as our first one has meant that a certain degree of unreliability has been associated with kicksorters using nickel wires.

This, of course, can be avoided by using quartz delay lines instead of nickel wires. This obviously makes a reliable device, but then you are back in the sort of money problem that you are in with the matrix store.

I think what Scarrott has told us this afternoon has shown that nickel lines are now extremely reliable devices and the work that he has done means that we now do know how to make them reproducibly. They will all do the same thing and the reliability of the device is at least as good as that of a magnetic memory.

Another possible argument is that a parallel analyser is still the faster in spite of the fact that any analyser is limited in speed by the period or length of the wave form used for comparison of pulse height. I think this is also a fallacy. It would only be true if the period associated with a serial memory were necessarily longer than that needed for comparison.

Accurate voltage comparison takes a time of the order of a microsecond in existing instruments. The mean comparison time for a 100 channel analyser is therefore of the order 50 to 100 μs. A singel quartz delay line to store, say, 1000 bits could have a repetition period of the same order as this. A cheaper wire delay of the same capacity would be considerably slower, but there is no reason why only one delay line should be used. The comparison is necessarily serial, and the memory should, I suggest, be serial, but there is no reason why the logical (adding) circuits should not be parallel.

If, for instance, you want to build an analyser in which you have a binary coded decimal display, you would naturally divide your bits into groups of 4; and if you use 4 delay lines, the logical processes of coding in these 4 lines could very well be done in parallel. If you do this and you require for instance 1000 bits, each of one microsecond, by using 4 lines, you reduce the lenth of the cycle period from 1000 to 250 μs straight away. This would imply a mean counting time of about 125 μs per pulse. There is no reason why this should not be taken further if in fact it is desirable beyond that stage.

The second thing I want to say about speed is that still the argument holds which was made when the original analogue digital kicksorter was brought out; that there is not very much advantage to be gained by going from, say, 500 μs to 100 μs in counting time.

The thing which might invalidate this argument, of course, is the advent of pulsed machines. The point is that although your counting rate is now not always random, in a large number of cases the pulse from the machine is short enough to make the number of counts which can be recorded from any one pulse using either an analyser of time 100 μs or of time 500 μs still of order unity.

If this is so, the advantage in going down by a factor of 5 is rather small. If pulses are coming in at random in time, then I still think that the time taken to collect a spectrum by an analyser

of dead time of the order of 500 μs is so small that it really is not worth considering in comparison with the subsequent data analysis time.

I should say perhaps that either type of analyser is equally amenable to all the sorts of convenience in display which reduce that data analysis time; such as for instance printing out on a pen-recorder or printing out digital numbers. In fact, I think both types are now available with both those facilities.

There is one point on future design with which I should like to end. Scarrott has described to us small logical elements for digital computers. The part of a kicksorter which digests the pulse after it has been compared is essentially just a small part of a digital computer, and as such its reliability and ease of manufacture can be increased in the same way which Scarrot has described, simply by making it of a lot of similar elements. If kicksorters are to be made in large quantities, then I think some thought should go into making all the logical parts of the circuits of very small units, in fact of logical elements of very much the same type as has been described.

NUCLEAR INSTRUMENTS 2 (1958) 117–120; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION II.5

ELECTRONIC INSTRUMENTATION IN NUCLEAR PULSE HEIGHT SPECTROSCOPY

BJÖRN ÅSTRÖM

Nobel Institute for Physics, Stockholm

At the Nobel Institute in Stockholm the decay schemes of isotopes in the lead region have been studied by several groups. In these investigations it was necessary to perform delayed coincidence measurements to determine the life-time of certain states in the nanosecond region. The decay schemes are often rather complicated and it is necessary to make some sort of energy selection when coincidences are measured. On the other hand, the sources available are rather weak and a pair of magnetic β-ray spectrometers with organic crystal detectors cannot be used for this type of measurements because of the low solid angles. We have therefore used NaI scintillation spectrometers and tried to improve the time resolution as far as possible[1]).

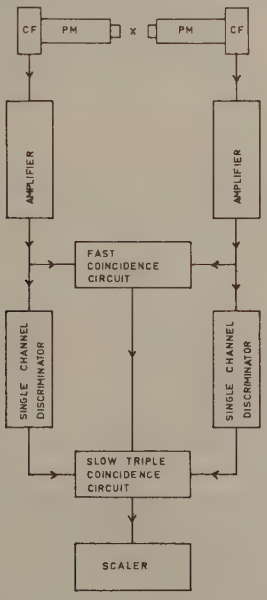

Fig. 1

The conventional arrangement is shown in fig. 1. Here we have two detectors, two linear amplifiers, a fast coincidence circuit, and a couple of pulse height analysers which select the proper energies in the two channels and are followed by a slow triple coincidence circuit.

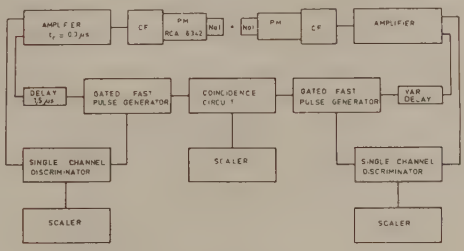

Fig. 2

The new electronic circuits used have been designed by mr. B. Johansson[2]) and should have applications also in other nanosecond measurements, for instance in time-of-flight experiments.

A block diagram of the basic arrangement is shown in fig. 2.

This circuit differs from the ordinary "fast-slow" coincidence arrangement by the fact that it only contains one coincidence circuit instead of the two circuits usually used. This makes it more easily convertable into a multichannel system measuring the complete coincidence resolution curve at the same time. This has been done by providing 25 coincidence circuits with their respective registers. The left hand fast

[1]) B. Åström, B. Johansson and I. Bergström, Ark. f. fysik 12 (1957) 205.

[2]) B. Johansson, Nuclear Instruments 1 (1957) 274.

pulse generator is feeding all the 25 coincidence circuits in parallel, while the right hand generator feeds a delay line with 25 successive taps each feeding one coincidence circuit.

The performance of the arrangement is demonstrated by the curves in fig. 3.

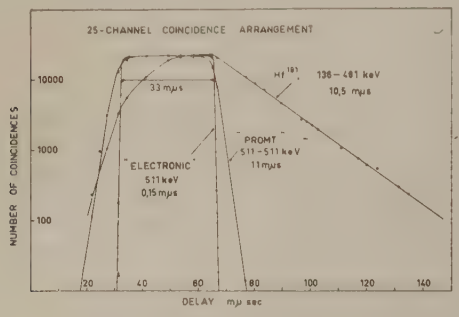

Fig. 3

The "electronic" curve was obtained by feeding pulses from one detector into both channels: this eliminates the time spread due to the scintillator and the photo-multiplier. The "prompt" curve was measured using the annihilation quanta from a Na^{22} sample.

According to Post and Schiff[3]) the slope of this curve can be calculated to be 0.8 ns for this case if the system responds to the first photoelectron in the photomultiplier. The slope found indicates that this is actually the case.

The third curve shows the well known delayed coincidences in the decay of Hf^{181}, for which the half-life has been measured by De Waard to be 10.6 ± 0.3 ns.

At the Nobel Institute pulse height spectroscopy is used for mainly two purposes namely to measure the energy of photons by means of scintillation detectors and to identify isotopes of the transuranium elements from the characteristic energy of α-particles emitted in their decay. In the last case an ionisation chamber is used as the detector. Some special features of the electronic instruments used in these experiments will be described here. The instruments as a whole are constructed following conventional practice[4]).

[3]) R. F. Post and L. I. Schiff, Phys. Rev. **80** (1950) 1113.
[4]) B. Åström, Ark. f. fysik **12** (1957) 215.

Gamma- and X-ray spectra are registered with a multichannel analyzer of the Hutchinson-Scarrot type. In order to present the data from this analyzer in a linear diagram a decoding device has been designed consisting of a keyboard with eight push-buttons connected to taps on a d.c.-voltage divider and a Brown Recorder. By pressing the buttons corresponding to the bright dots on the CRT screen, a d.c. voltage is developed, which is proportional on a linear scale to the number of counts registered in a certain channel. This d.c.-voltage is measured by the recorder. In this way it is possible to have a spectrum transferred from a photographic picture of the CRT screen into a histogram on a linear scale in about five minutes.

On other occasions γ-spectra are recorded with an automatic single channel spectrometer[4]). The block diagram is shown in fig. 4.

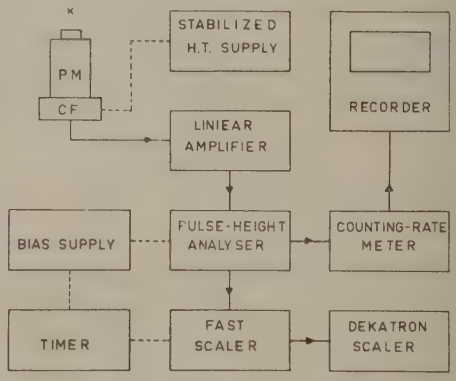

Fig. 4

By using a channel width that varies with the channel setting[5]) it is possible to record spectra containing both low and high energy peaks without changing the sensitivity of the counting-rate meter or the recorder. Fig. 5 shows the difference in spectrum shape when a 1 volt and a 2% channel, respectively, is used.

For the measurement of α-energies an ion chamber and a 50 channel pulseheight analyzer of the "stacked discriminator" type is used[4]). The block diagram is shown in fig. 6.

The pulses in each channel are counted by mechanical registers of the post office type,

[5]) B. Åström, Nuclear Instruments **1** (1957) 143.

which determine the time resolution of the instrument. To trigger a certain channel it is necessary to apply both the pulse to be measured

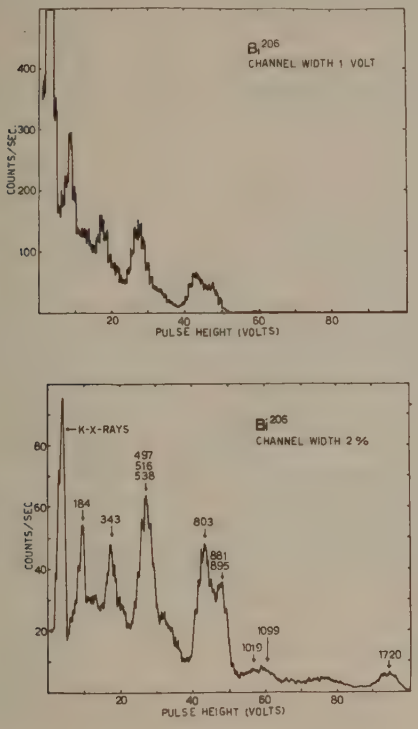

Fig. 5

and an inspection pulse, which is common to all channels. The insertion of one scaler in the

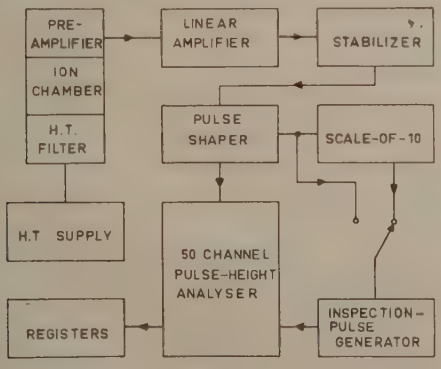

Fig. 6

inspection pulse path has thus essentially the same effect as a scaler in each channel. The randomness in pulse distribution is smoothed out and the pulsehandling capacity in each channel becomes much higher.

The stabilizer has been incorporated to improve the over-all stability of the spectrometer when long runs are made. It consists of a motordriven attenuator controlled by the counting rates in two channels placed on different flanks of a reference peak in the α-spectrum. It keeps these two counting rates equal and thus locks all α-peaks to certain channels, independent of what causes a drift of pulse amplitude in the system. The pulses in the stabilizer channels are counted by dekatron tubes and when 10 or 100 pulses are accumulated in one channel the counting in that channel stops and the motor starts to change the amplification towards a higher counting rate in the other channel. When this channel has received the same number of pulses the motor stops and the cycle starts again.

The reference line for the stabilizer is obtained from a Po^{208} source ($E_\alpha = 5.10$ MeV) where

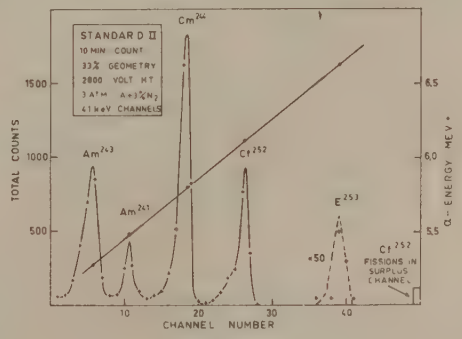

Fig. 7

the energy of the α-particles has been decreased below 5 MeV by absorption. The reason for using this type of reference sample is that one wants an α-peak with a fairly high counting rate (about 200 c/min) but with an energy lower than any energy that one might want to measure. Still, some counts from the reference source are obtained in higher channels because of pile-up effects. Therefore, in measurements where one

wants the lowest possible background, this source must be taken out of the chamber and the stabilizer cannot be used.

The spectrum of a standard sample is shown in fig. 7. The full width at half maximum of the Cm^{244} peak is 55 keV.

The instruments described here have partly been put together from smaller standard devices such as linear amplifiers, scalers and so on, which are built as complete units with their own power supplies etc.[4]). A special mechanical lay-out is used which facilitates maintenance and testing and improves the cooling by concentrating all tubes in a well ventilated closed space where they can not heat other components.

DISCUSSION

MAEDER: The effect of the constant relative channel width display has been very convincing in your example; does one not always use this feature, e.g. with multichannel equipment. This would merely amount to the introduction of a logarithmic pulse height to time converter, which could be used with any multichannel system. In this way the complete spectrum could be covered with a logarithmic energy scale.

HUTCHINSON: In a few months, there will be a commercial analyser available with this feature.

ÅSTRÖM: I have only been talking about a single channel analyser. We have not yet tried this for multichannel arrangements.

WELLS: What is the delay between channels in the 25-channel analyser for which you have to compensate?

ÅSTRÖM: I use a 1.5 μs lumped parameter delayline.

WELLS: What transient response does it have? You are dealing with times of the order of ns, so you must preserve the rise time of the pulses.

ÅSTRÖM: The rise time of the linear amplifier is about 0.3 μs. The rise time of the original pulse need not be preserved because a new fast pulse is generated which is time-corrected for variations in pulse-height.

NUCLEAR INSTRUMENTS 2 (1958) 121–127; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION II.6

DECIMAL MEMORY SYSTEMS

D. MAEDER

Eidgenossische Technische Hochschule Zürich, Switzerland

The inconveniences of the binary notation are well known to all users of those pulse spectrometers which have no other display available but the binary one. Modern storage type spectrometers usually have one or more artifices to facilitate the reading of the data; for instance, they may be equipped with:

(1) *Analog display* on a CR tube, which suffers the loss in accuracy common to all analog displays of numerical data;

(2) *Neon lamp indicators* which present one channel at a time, with manual switching from channel to channel. This type of decimal indication is usually intended only for routine tests of the memory operation;

(3) *A mechanical print-out system*, using adding machine tape;

(4) *A mechanical curve plotter* based on a servomechanism.

That these mechanical expedients are not supplied with all storage type spectrometers may be taken as an indication that even the manufacturers of these complicated instruments consider the mechanical printing or plotting devices as some sort of luxury. The following two arguments will strengthen this opinion:

(a) Although the mechanical devices are quite expensive, they are much too slow to follow half-lives of the order of a few seconds.

(b) They do not save the experimenter as much time as it might first appear, because he still must perform arithmetic operations on hundreds of numbers which must be read individually from the prints.

On the other hand, if one would like to feed the record directly into a computer, then the decimal print out is certainly not needed. However, there exists the possibility of operating the memory directly in a decimal manner, with some waste of storage capacity. This waste must be balanced against the saving of equipment achieved by the elimination of the need for a binary-to-decimal converter mechanism.

Decimal storage is possible with all known types of storage circuits. In the magnetic core memories, for instance, one can use a straight binary code (requiring 3.3 bits per decade), or a hexadecimal code[1]) (requiring 4 bits per decade) which compromises storage capacity for easy reading. In a commercial version of such an instrument (PA4, manufactured by PENCO) a mixed storage system is used in such a manner that the less significant places are represented in binary form and the more significant ones in coded decimal form. It is important to note that in a magnetic core type memory, every bit corresponds to an actual circuit element. This fact is in contrast to the storage characteristics of either the delay-line or the Williams tube memories, which lend themselves particularly well to decimal modifications.

DELAY-LINE STORAGE

McKibben *et al.*[2]) at Los Alamos have developed a quartz plate delay-line spectrometer which works in a hexadecimal system (4 bits per decade), giving a coded decimal display. The authors state that an unskilled person can read 100 channels in 15 minutes. Evidently the relative economy of bits in that apparatus still seems to be paid for by some waste of time of the human operator.

[1]) P. W. Byington and C. W. Johnstone, IRE Convention Record part 10 (1955) p. 204.

[2]) J. L. McKibben, J. D. Gallagher, and H. J. Lang, IRE Convention Record part 10 (1955) p. 186.

With the decimal delay-line apparatus which was developed at Zurich[3]), we have been much more generous with bits, allowing 11 bits per decade. The display of the memory contents is correspondingly more convenient to read. Fig. 1 shows the scintillation spectrum produced by γ-rays from Hg^{203} (278 keV + X-rays of about 72 keV).

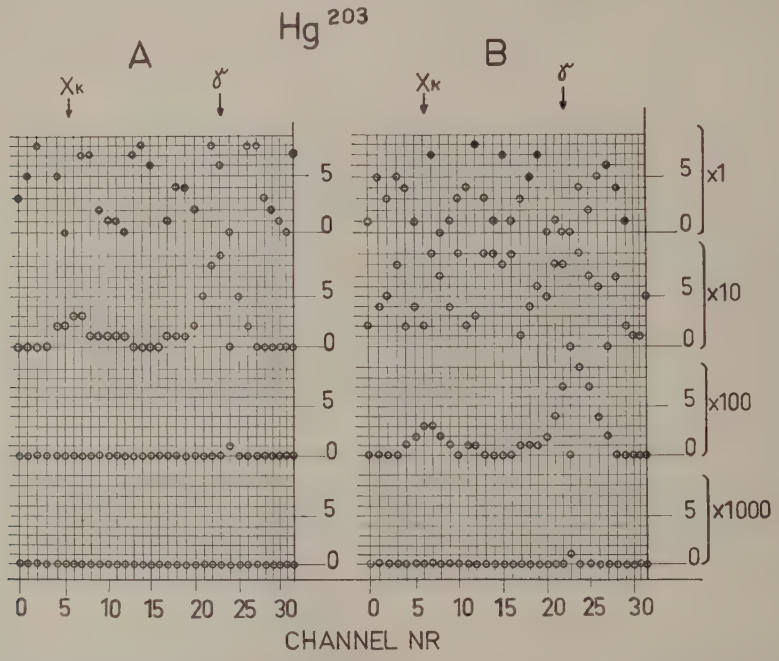

Fig. 1. Display of delay-line memory. Automatic recording after a score of (A) 100 counts, (B) 1000 counts in one channel.

The apparatus can be made to stop at some preset count (for instance, when the channel which counts at the highest speed arrives at 1 000) at which point a photograph is taken automatically.

This type of decimal storage combines an analog display (with a linear indication of the spectrum shape) with the full accuracy of a digital storage system.

The organization of the decimal adding and subtracting processes is explained in fig. 2. The memory uses 10 bits per decade, separated by one dead interval which is identified by the "decade marker pulses" generated in the timing circuitry. The memory line contains always one pulse per decade. At the beginning of the measurement the pulses are the "0" position. If "1" is to be added, then the first "0" in the particular channel must be shifted by one digit. The next addition gives another shift by one digit, and so on. If there was a nine, the shift will produce a pulse in coincidence with the succeeding decade marker pulse. The coincidence is fed to a circuit which creates a new pulse at the "0" position 10 bits in advance and at the same time causes the shifting operation to continue. If the original number was 99, another coincidence will occur with the second decade marker, and the shifting will continue until there are no further coincidences between "9 + 1" and decade markers. We can of course make such a system work in the reverse direction as well; and this was actually the case when the apparatus was used for taking the pictures shown in fig. 1, as is evident from the inversed order of the decimal places.

Fig. 3 shows the circuit arrangement for the

[3]) D. Maeder, Helv. Phys. Acta 29 (1956) 459.

adding operations. When the carry operation occurs, we have to create a new "0" 10 bits in advance. In order to avoid the use of electromagnetic delay lines for producing such a large advance we prefer to put an auxiliary transmitter at the appropriate position on the magnetostrictive memory line to write the advanced "0" at the instant of the "9 + 1" coincidence.

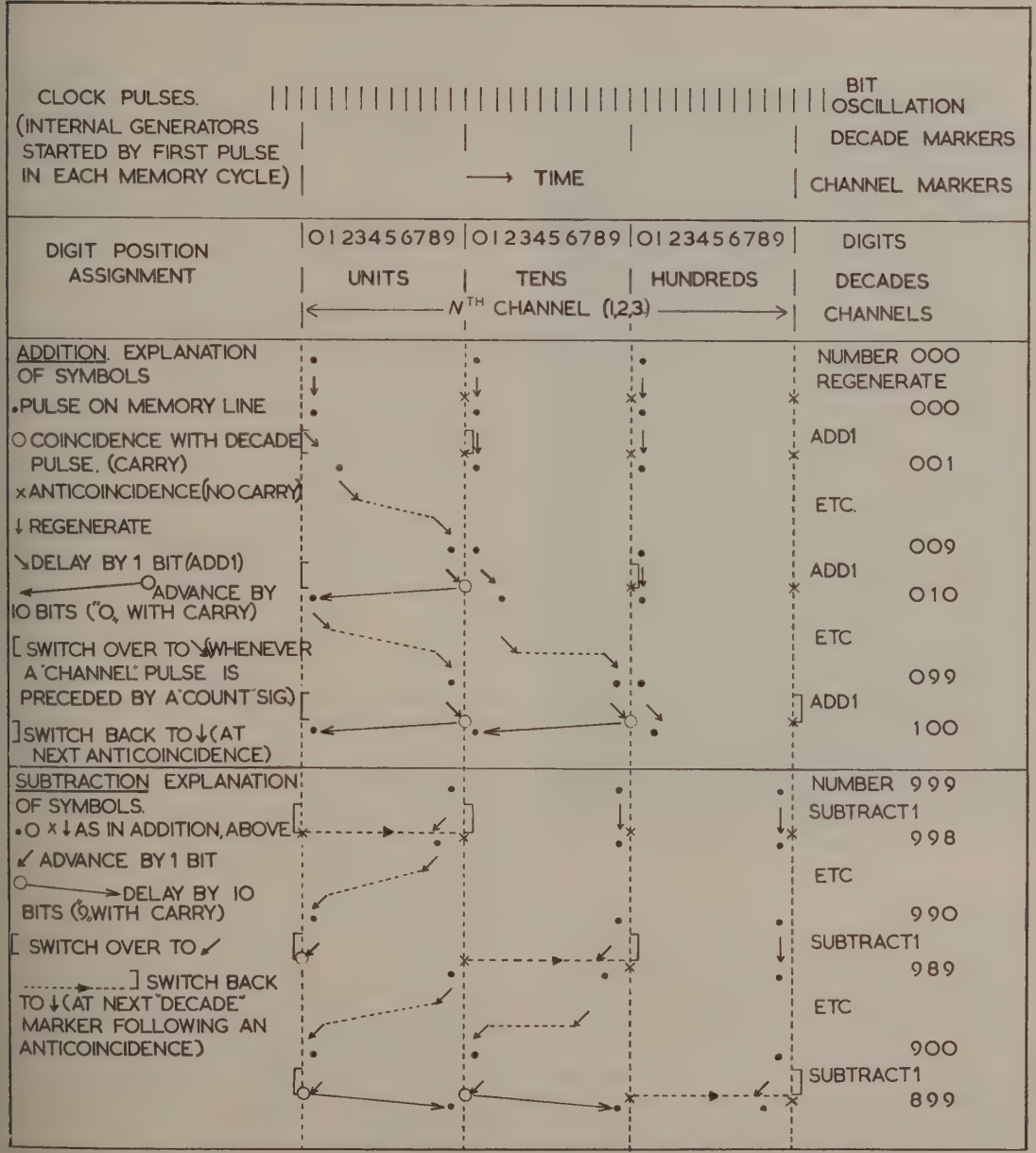

Fig. 2. Block diagram of a proposed dots-on-film recorder. (Quantized p.a. deflection, semiquantized time axis).

Fig. 4 is a block diagram for the reverse operation circuit which we actually used in

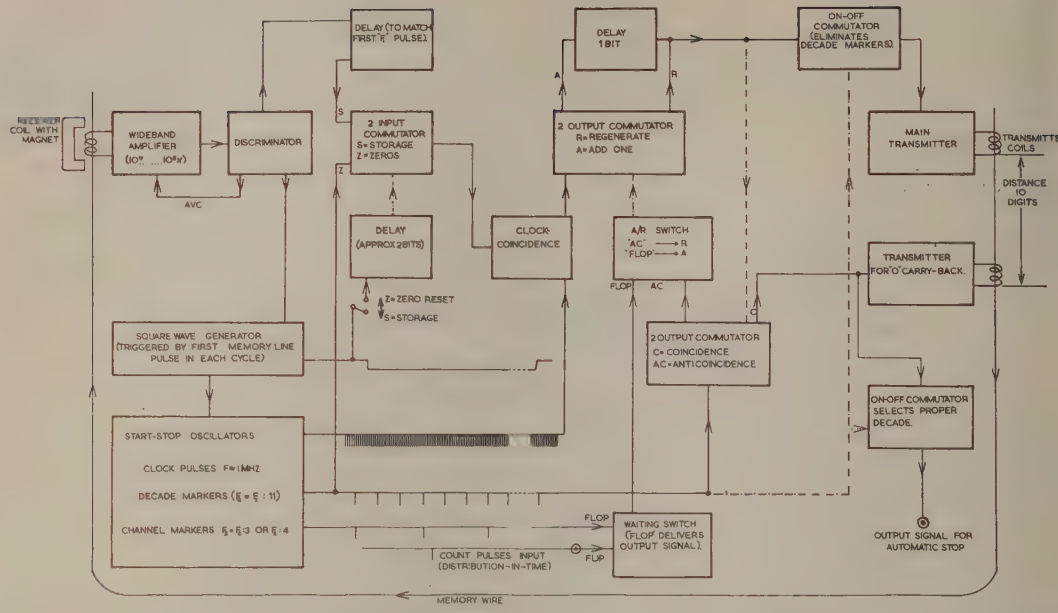

Fig. 3. Block diagram of decimal memory circuit used for adding time-distributed counts. A "commutator" accepts either of two inputs (or feeds one input to either of two outputs) depending on the signal introduced by the dot-dashed line.

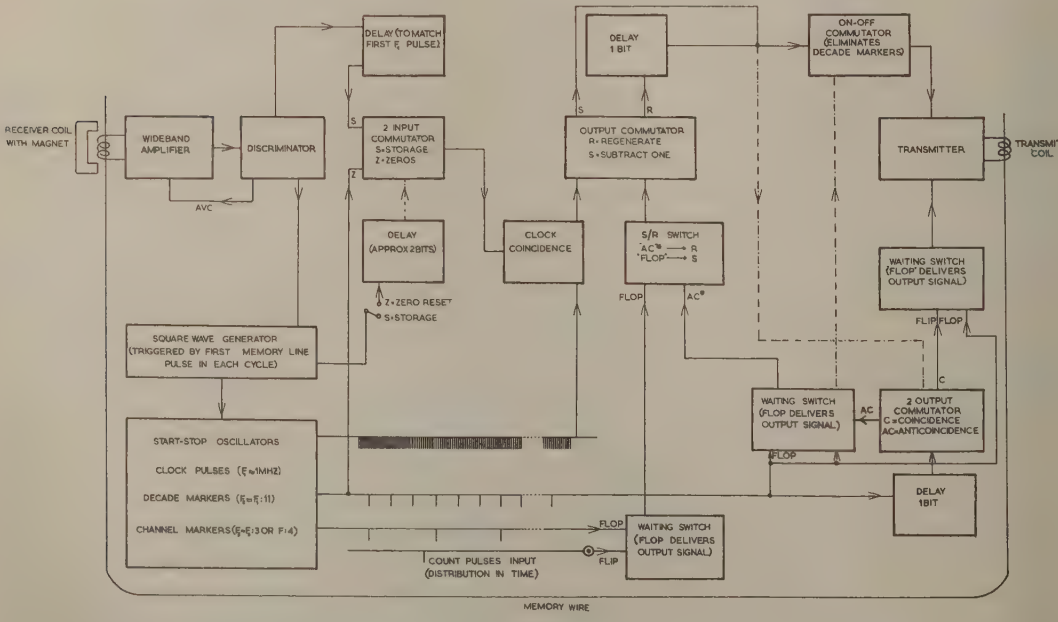

Fig. 4. Block diagram of decimal memory circuit used for subtracting time-distributed counts. The "automatic stop" signal output is not shown.

making fig. 1. It is very similar to fig. 3, except that (1) the order of the delay between the "regenerate" and the "subtract" or "add" positions is reversed, and (2) the auxiliary transmitter is replaced by a flip-flop circuit which produces the required 10-bit-delayed output signal by waiting for the next "flop"-input pulse whenever it had been prepared to do so by a preceding "flip"-input event. (This is the basic operation of all "waiting circuits" contained in our diagrams.) The delayed "carry" pulses are then handled by the main transmitter. Both of the schemes shown in figs. 3 and 4 were tried out because we want to combine adding and subtracting in an improved version of the spectrometer.

Although we must use 11 bits per decade, the storage capacity of a given memory line is not necessarily reduced by a factor 11:3.3, since the pulses on the line are always separated by at least one empty bit, and thus can never follow each other as closely as in a binary system. Therefore, it should be possible to operate a given memory line with a decimal adding network at twice the normal bit frequency, so that the loss figure reduces to 5.5:3.3. A further increase of the empty interval between decades may be expected to improve the situation still more.

On the other hand, the straight decimal system has two definite advantages over both the ordinary and the modified binary systems:

(1) There is no AVC problem arising from variable filling of the memory, since the total number of pulses on the line does not change as the information comes in.

(2) Since exactly one pulse must be on the line per decade, errors will be recognized very easily, since the vanishing of the pulse in a decade or the appearance of two pulses in a decade is easily observed.

ELECTROSTATIC MEMORY

We have based our development of a decimal electrostatic memory on a suggestion by Kilburn and Ord[4]). These authors explained the principles of a decimal system, without giving circuit details, and, in addition, they reported some preliminary experiments. We have tried to build an apparatus of this kind a the ETH.

Fig. 5 illustrates 8 channels (each containing

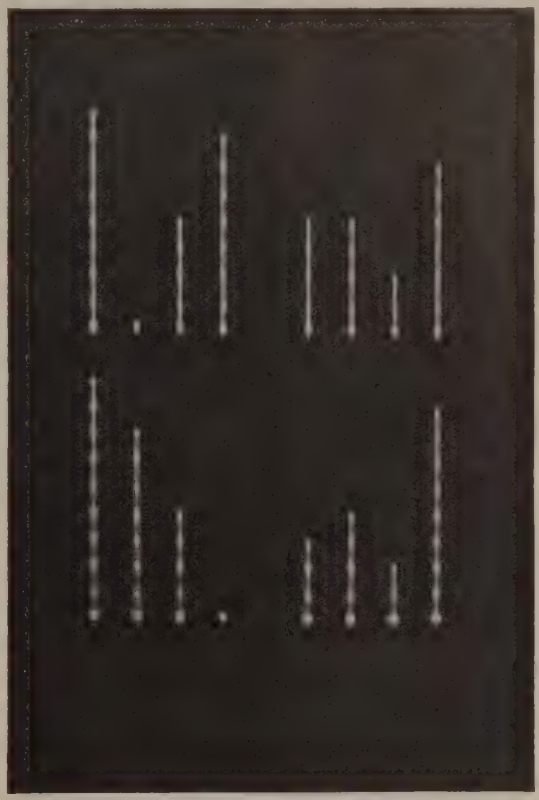

Fig. 5. Display of electrostatic memory.

2 decades), stored on the screen of a 6571 Williams tube. For example, the first channel to the left shows the number 89. The regeneration of the information normally starts at the "0" position. If there was already a 0 (as in the fourth channel), then there is a charge left on the screen from previous scanning, and no signal of appreciable amplitude will be produced at the pickup electrode. Absence of a signal immediately stops the scanning for the particular decade, and the beam will then be shifted to scan the next decade.

On the other hand, if there was a number = 0 written in the channel under consideration,

[4]) Proc. IRE (II) 100 (1953) 513.

then the scanning at the "0" position (where no charge will be found now) will produce a positive signal at the pickup electrode. This signal will cause the beam to displace to successive number positions 1, 2, 3, ... until a negative signal shows that the place containing the stored charge is hit. In this position the beam is switched off, leaving the charge at its previous place.

The operations involved in the regeneration,

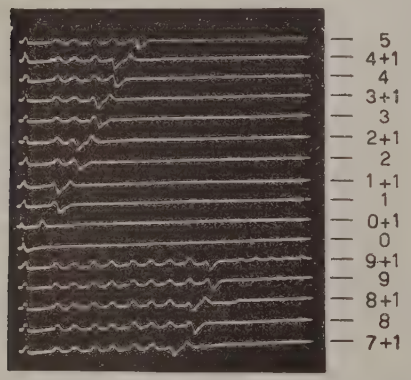

Fig. 6. Signal picked off from screen of electrostatic storage tube in a decimal memory system. Every other sweep (beginning at the bottom of the figure) shows the adding of one unit.

as well as in an "add 1" will be seen from fig. 6. Each successive sweep in the figure displays the signal generated at the pickup electrode for a different charge pattern, produced alternatively by the "add 1" process, or by the regeneration of the charge distribution, beginning with the regeneration of the pattern representing "8" (second sweep), then showing "8 + 1", "9", "9 + 1", "0", "0 + 1", etc. in succession. The 10-step quantization of beam positions is produced by a 6700 (M0 10) magnetic beam switching tube whose targets are connected via diodes to the 6571 tube deflection plates. Since the 6700 tube operates in a cyclic manner, the "carry" operations require no complex circuitry. The flyback from "9" to "0" in the 6571 tube must be slowed sufficiently to allow the charge distribution to follow the scanning beam.

In the present state of our apparatus, the scanning of a full decimal requires about 60 μs if it stores a "9" and four μs if it stores a "0". A "carry" requires another 40 μs. Thus, if a channel is counting at full speed, it will occupy on the average about 36 μs per decade. According to preliminary data published by Kilburn and Ord one may expect that it will be possible to reduce this time to about half of the present value. With 4 decades per channel, and an arrangement such that on the average only half of the full dead time is effective when the pulse spectrometer accepts statistical pulses, the dead-time contribution caused by the electrostatic storage would at most amount to $18 \times 4 \times \frac{1}{2} = 36$ μs, which compares favorably with dead times in serial memory and even in some magnetic-core memory spectrometers.

DISCUSSION

HUTCHINSON. For taking the data out of a kicksorter, you use a system in which you can have a digital display which is at the same time very nearly analog. For myself, I have been very happy with a binary system, because octal numbers are quite easy to use, and in England (at least) we have graph paper with one inch and one eighth inch squares, so you can straight away use octal numbers. But if you are going to use kicksorters in large numbers, and if you expect to use unskilled operators then you will not wish such operators to use this system. Firstly you need an analog display which shows you at once if the peak is at the right place, or if you ought to shift your bias. You should not ask this display to give any accurate information at all, so that it could be looked at by an unskilled operator. And secondly when the experiment is assumed to be at a suitable point, you want to take down the information in a printed or punched digital form. That means that you want really to take it down in some mechanical way and not to rely at all on the unskilled observer.

MAEDER. This takes 2 minutes; what are you doing then for measuring half-life of 10 seconds?

HUTCHINSON. This is rather a special case. If you are going to have an unskilled operator, then you do not want that operator taking down digital information from a photograph or from a screen. So at this stage I think you want to have the things printed out. You can make the print-out in decimal by using the same sort of binary coded decimal that you described from Los Alamos. This

is not difficult. You can also make the thing completely analog. Do you not think it is better to have these two things available (which you could get in turning a switch) than to lose the information storage-capacity which you do by using 11 bits per decade?

MAEDER. As I already mentioned, the comparison between 11 bits and 3.3 bits used per decade is not quite fair, because we can make a trade with the empty bit between successive decades. We might even go one step further and use two empty bits between decades, so that the closest spacing of two pulses on the memory line (namely, a "0" and a "9" in two adjacent decades) would become 3 bit periods. With a given memory line, we could then use triple the maximum bit frequency at which it would operate in a binary system, so that the actual time used per decade would be $12:3 = 4$ as compared to 3.3 for the ordinary binary system. Thus I believe that such a decimal memory, which gives at once the semi-analog display and the exact decimal indication without using any mechanical parts, does not necessarily loose an appreciable fraction of the storage capacity available with binary systems. However, all kinds of CR tube displays might become obsolete some day because one would like to obtain a record which fits directly into the computer; besides, neither the printed tape nor most of the other presently available recording systems are suitable for this purpose.

HUTCHINSON. If you are thinking of the two separate uses of a kicksorter, is it not then better to have two separate displays available: a binary display which is simply what you are using for a computer and a second decimal display?

MAEDER. Yes, if you can afford to have both.

KANDIAH. I have a feeling that this question of taking a very quick record down is somewhat overworked. I have gone through the arguments about phenomena with a short half-life, and how you record information in these cases, designing in the devices which we made in Harwell. The information is presented there in the following forms: There is a continuous analog display with a repetition rate of greater than 10 s^{-1} which will not have too much flicker. The absolute accuracy of the analog display is only 5%. But the relative accuracy between any two channels on one sweep of the display, which takes 4 ms, is better than 0.2%, because the same circuit is looking at every channel in turn. When you are actually recording this information quickly say in two minutes, because you are dealing with a short half-life of only a minute the maximum number of counts with a dead time of 1 ms is only about 100 000 counts. The information you collect in that time is not sufficient statistically for a measurement of the order of 0.1%. In fact, a simple photograph of the analog display once every 10 s will give sufficient accuracy and you have to add nothing more to the device. We have decided that for the persons who want this short half-life measurement, we will do this, and for people who want to read out accurately in decimal digits, then they have to be prepared to spend either a quarter of a second for each channel which is the time for the device to convert into the decimal indicator, or a quarter or half a second per channel to go into an automatic record on a tape for use with a computer. But there is actually another solution to this which we have been considering. In the system of automatic display at every 1/20th interval it is possible to arrange the logic of this display so that instead of spending only 4 ms it spends a little longer every few seconds, and during that time, it puts the digital information bit by bit into tape, so that the tape takes down the digital information. The tape can then be taken away and put to the computer.

MAEDER. Does the relative accuracy which you mentioned to hold for the analog display apply also with two peaks which are in different decades?

KANDIAH. The thing to be remembered of course is: if you are interested in two peaks which occupy say decade A and two decades down for example, then no linear analog display will be adequate because you are limited by the size of the spot. An alternative would be to have non linear display which is what we don't desire; the real solution is an analog display whose range can be switched since one would then have a smoother curve than that in which different decades are displayed separately.

MAEDER. You need more than one switching if you have to show all decades, whereas with the decimal storage you need no switching at all to show up all the information.

KANDIAH. Except that it takes you a little time to operate the switch on each decade, there is not much complexity. If you have a pure analog display as you suggest, the time taken to shift one's eye from one decade to the other would be just as long as to turn a switch.

VON DARDEL. I would like to ask question if some of you has an experience on an idea, which we would want to try, namely instead of having a complicated pulse-height analyzer with special printout, to photograph the digital display of the pulse-height analyzer, and them make a special scanning device which takes the information from the photograph and transforms it in a more convenient form, say analog or decimal or whatever one likes. In this way, the printout would not waste the time of the pulse-height analyzer which could be started again as soon as the photograph has been taken. Of course this is rather a big problem.

HUTCHINSON. I do know how much this short half-life problem is really going to come up. Would it not be possible that one way of tackling this is really recording the pulses on tape. Some resolution is lost in this process but if one is really after a short half-life this may be a small price to pay. Once you have the information on tape, both time and pulse height of every pulse, you can derive any sort of information you want from it.

NUCLEAR INSTRUMENTS 2 (1958) 128–129; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION II.7

HIGH SPEED TRANSISTORIZED SCALE-OF-TWO

E. BALDINGER and P. SANTCHI

Basel

Fig. 1 shows a high speed scale-of-two circuit. With Philco transistors SB-100 we achieved a resolution better than 50 ns for paired pulses. The principle of operation of this circuit may be explained as follows: Transistors no 1 and 2 are DC-coupled and the circuit has therefore two stable positions (no 1 conducting or nonconducting; no 2 nonconducting or conducting respectively). The hole storage effect in the transistors 1* and 2* is used as a short time memory, which remembers the position of the circuit just before triggering. If we assume transistor 1 to be conducting the associated transistor 1* conducts

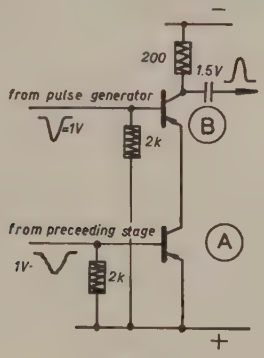

Fig. 1

a corresponding collector current. The bias conditions are such, that its emitter-collector voltage is a few tenth of a volt. Transistor 1 is non saturated. Since transistor 2 is cut-off the transistor 2* has no collector current and is therefore saturated, and holes and electrons are stored in the base region. A positive input signal applied simultaneously to the bases of 1* and 2* will immediatly cut off the collector current of transistors 1 and 1*. Due to the holes stored in the base region of transistor 2* it is still able to deliver a collector current and the cut off signal at the base electrode will even help to produce such a current. It is thus evident, that the input

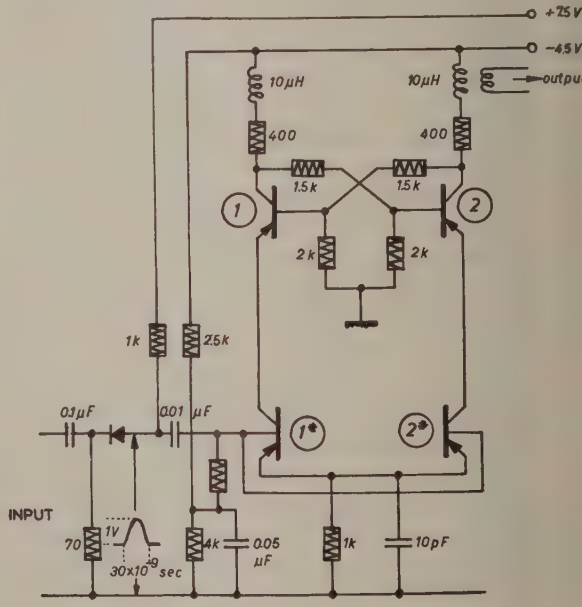

Fig. 2

signal will cut off the previously conducting side of the circuit and will help to impress a current on the other side. This action permits a high speed switching of the scale-of-two. The two compensating coils in fig. 1 will further increase the switching speed. The signal for the next stage of the scaler is taken from one of these coils. Unfortunately it is not possible to load the output of the first stage of the scaler with the input impedance of the next one. An amplifier has to be inserted. Since it is rather difficult to amplify pulses of about 30 ns time duration with Philco SB-100 transistors, a 3-stage tube amplifier

(supply voltage: 140 volts, total current 20 mA) was used in the first experiments. The problem of amplification may be solved partially by using a suitable transistorized gate circuit as shown in fig. 2. In this gate circuit the larger fraction of the output energy is delivered from the input signal to the first scale-of-two. The output of the first scaler is used only as a gate signal applied to the lower transistor A in fig. 2. This signal will introduce holes in the base region of transistor A. By applying a delayed strong signal to the upper transistor B these holes can be drawn away rather quickly to the collector. By this integrating action of the base region of transistor A, the peak value of the output current of the gate may be 3 to 6 times greater than the peak current in the base lead of transistor B.

To couple the second to the third stage of the scaler a gate circuit is no longer necessary and a coupling transistor has sufficient amplification.

NUCLEAR INSTRUMENTS 2 (1958) 130–135; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION III.1

INFORMATION HANDLING SYSTEMS FOR NUCLEAR MEASUREMENTS

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Problems related to evaluation of information recorded during a nuclear experiment are intimately connected with the manner of presentation of the data to the observer. Paradoxically, the most flexible apparatus is generally the most tedious with which to work. The observer has to undertake total analysis of the data as recorded on the film but has before him all details such as variation in pulse height, statistical behaviour, zero level stability, unwanted effects due to disturbances and so on.

Today, specialized equipment is incorporated

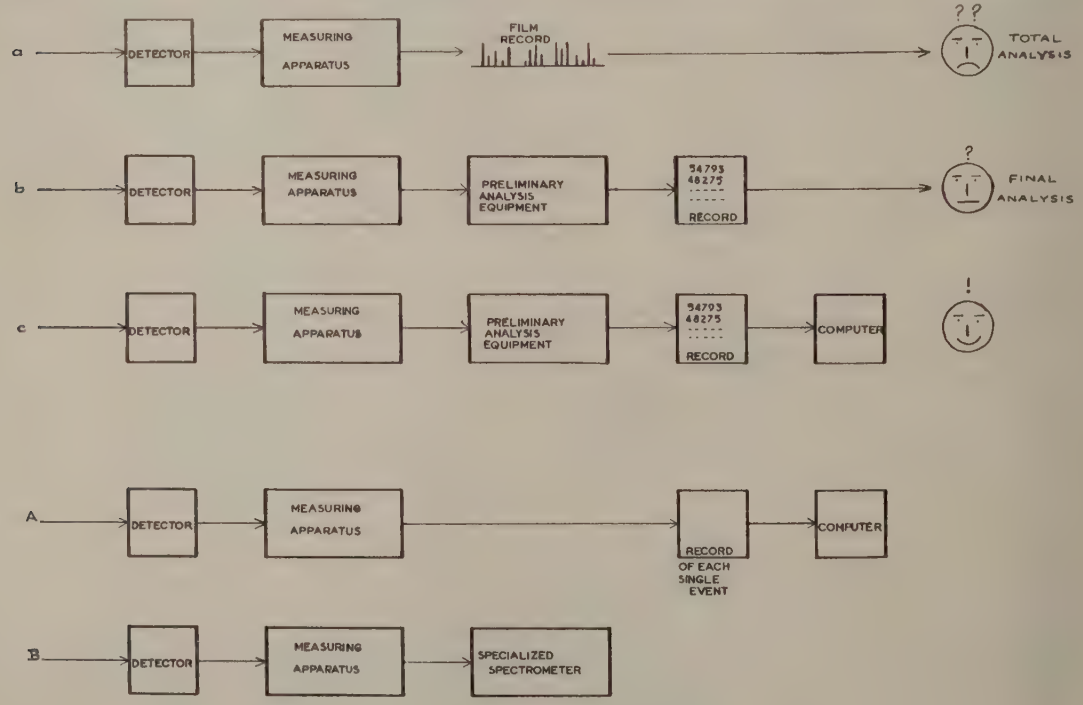

Fig. 1

For example, in the classical form of a simple counting experiment, the radiation under study is examined by means of a detector coupled to a simple measuring apparatus displaying pulse heights which are recorded on a moving film. to perform some preliminary analysis on the databe fore recording. For example, in the simple counting experiment, a scaler would be used to give the total count at the end of a run. This is a more convenient arrangement since the observer is now only required to reduce the final data.

† Now at Oak Ridge National Laboratory, Tennessee.

However, the use of the scaler alone deprives him of detailed information on individual pulses. In order to retain this detail more complex apparatus is needed.

The future trend toward more complicated experiments will eventually require the use of a computer which carries out all subsequent analysis. Thus minimum effort will be called for on the part of the observer who can devote his time to planning new experiments. Usually computers will not accept data occuring at random times; some intermediate storage will therefore be necessary for smoothing.

Fig. 1 depicts the evolution of the experimental arrangement. If unnecessary parts are eliminated, two simplified cases arise:

(A) Each event is recorded individually without preliminary analysis in such a manner that all associated information is retained in the record which is then fed to a large conventional computer for analysis.

(B) Instead of using a separate computer, the measuring apparatus feeds directly to a highly specialized spectrometer which performs all the analysis.

It is evident that case (A) will lead to the more flexible equipment. Various known or conceivable approaches to these schemes will now be discussed.

1. Case A: Recording of each Individual Event

The computer must be fed from a coded record, whereas many methods of data recording give an analogue presentation. Such records may however be fed into an automatic scanning device for conversion into coded form. It is therefore profitable to first consider the various possibilities of *analogue recording*.

(1) *Photographic recording* of pulse height distributions. In the classical form, fig. 2a, the whole pulse from zero to peak value illuminates the CRT screen so that nearly coincident events (e.g. the fourth and fifth) are difficult to resolve into their proper amplitude channels. By so arranging that only the peak of the pulse illuminates the CRT screen (giving a dot display, see fig. 2b), this difficulty can be avoided except where nearly coincident pulses have almost the same amplitude. This method of dot recording, which conserves the original pulse amplitude and time distribution, permits the use of 100 channels with a channel width stability of 2% over the spectrum. The film can be photoelectrically scanned. However, at a film speed

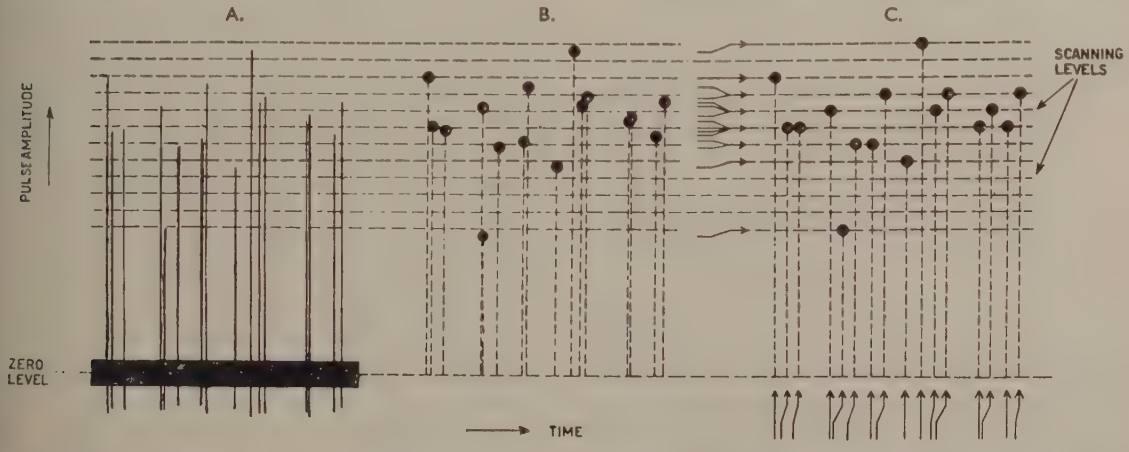

Fig. 2. Appearence of the same set of input pulses when recorded on moving film by: A. Classical method (continuous illumination) B. Proportional dot recording system (conserves original P.A. and time interval distributions) C. Quantized dot recording system (proposed in order to remove ambiguities in evaluation).

of 50 cm/s, the overlap losses will exceed 3 % in all channels counting at > 100 pulses/s. This method has been used at Iowa State College, M.I.T. and Case Institute.

A further improvement can be made by quantizing the pulses in amplitude and time. Amplitude quantization can be effected, for example, by stepwise discharge of a capacitor, and time quantization by applying a similar

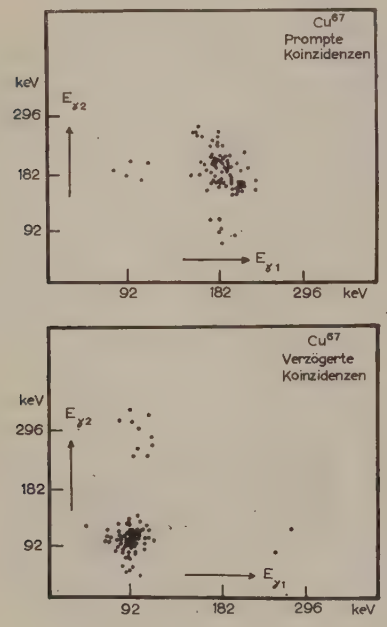

Fig. 3

technique to the CRT deflection paralleling the film transport. In this way (fig. 2c) both ambiguities can be removed and pulses accepted at a rate up to 20 000/s (compatible with the fastest film speed available) without any overlap.

(2) *Printing on a moving chart* is very slow and the record cannot easily be scanned photoelectrically.

(3) *Magnetic tape recording*. Here the pulse height is represented by the intensity of magnetization of the tape. The output signal is thus a measure of the orginal pulse height. According to Cavanagh (Harwell), this correspondence is accurate and reproducible to about 3 %. Overlap losses are a serious limitation, and time quantization would not appear to be feasible. At a tape speed of 2 m/s and a counting rate of 200/s, the overlap loss is about 1 %, but the tape would be 99 % empty. One very strong advantage af tape, especially for coincidence work, is the ability to have several tracks per tape. Thus, pulse heights could be recorded simultaneously from many inputs. With one-inch tape sixteen tracks are possible.

(4) *Photographic analogue of tape.* If the photographic density of a dot on the CRT screen is used as a measure of pulse height instead of using dots of constant intensity, the film could be amployed in a manner analogous to magnetic tape except that 100 independent tracks would be available. This method is rather inaccurate—at best a reproducibility of ± 10 % in pulse height might be expected. Therefore, only about ten levels of pulse height could be assigned to each track. On the other hand it would be possible to fill about 90 % of the film area with data.

A special case of dot density recording arose from experiments similar to those shown in fig. 3, which represents two-dimensional recording of prompt and delayed $\gamma - \gamma$ coincidences from the decay of Cu^{67}. Approximately eighty events are depicted, but counting the dots individually is a tedious task. With a large number of events for good statistics, the task becomes impossible but the resulting smear can be analysed by plotting lines of equal photographic density. This method has been used for the two-dimensional recording of $\beta - \gamma$ coincidences at the E.T.H.

Coded recording, as previously mentioned, is the only suitable method for directly feeding a computer. The information is recorded in discrete steps e.g. if there are n "bits" each with a "yes" or "no" possibility, $2^n - 1$ values or channels of the input variable can be recorded in binary form. It is necessary to group the n bits together in some logical fashion such as punching n holes in a row. The recording medium will now be considered.

(1) *Punched cards* with rows of holes are slow to operate.

(2) *Punched tape* with seven holes in a row allows at most 127 channels to be coded into one row. Normal tape speeds are 20 rows/s, but with special machines this can be increased to 60/s. A few per cent loss is encountered at a statistical rate of 1 count/s. Multidimensional information can be punched on tape by having several independent inputs operating in time sequence, as has been done by Detenbeck (Princeton) and Hermannsfeldt (University of Illinois, Urbana) but this clearly increases the dead-time in proportion to the number of inputs.

(3) *Magnetic tape* with eight tracks would allow the use of 255 channels. The tape introduces about 1 % loss at 200 counts/s. One-inch tape with sixteen tracks would permit two independent inputs each of 255 channels or three independent inputs each of 31 channels so that subsequently two- or three-dimensional distributions could be analysed. Multidimensional information could be recorded by adopting the time sequence technique referred to above.

(4) *Photographic recording*. The large capacity of 100 tracks could be divided in many different ways e.g. fourteen inputs sorted into 127 channels each. Recording speed would surpass that of all other methods, reaching 20 000/s with megligible loss if time quantization is employed.

2. Case B: Use of a Specialized Spectrometer

In order to investigate the other type of arrangement, in which the spectrometer itself is used as a computer for automatic evaluation of the data, it is necessary to consider the type of computation encountered. This discussion will be limited to one dimensional analysis since such an apparatus will be somewhat specialized.

The simplest case is the subtraction of the background distribution from the measured pulse amplitude spectrum. A more complicated case, which can be resolved into a number of multiplications and subtractions is the analysis of a complex pulse height distribution into its components. Each component has a complicated shape which must be determined by separate calibration measurements or by means of auxiliary computations. If the shapes of individual components are known as a function of energy, linear combinations thereof have to be subtracted from the complex spectrum until nothing remains. Thus a large number of multiplications and subtractions are required with the coefficients adjusted to give a good fit—usually judged by visual inspection of the remaining distribution.

This type of calculation could be performed by the spectrometer itself and fig. 4 shows a possible arrangement in which a spectrum photographed from the CRT display of a Hutchinson-Scarrott instrument may be read back into the memory after reversal processing of the film. Briefly the operation is as follows:

With the operation switch (M/R) in position R, pushing the press-button releases the intensification, in the CRT, of the first digit in each channel in the course of the next memory cycle. If a particular channel carried a *one* when the original photograph was taken, the film positive will transmit the corresponding light pulse to a phototube, which tells the memory circuit to add *one* (or substract *one*) at the appropriate place in that cycle, i.e. to change *zeros* into *ones* or vice-versa until the first *one* appears either at the output (ADD position) or at the input (SUB position) of the reproduce-or-change network. In subsequent cycles the other digits are intensified in turn, until the scanning terminates when the cycle scaler returns to its zero position. The same process could be repeated in order to multiply the scanned spectrum by any integral number. A more rapid multiplication procedure is indicated in fig. 4 with a switch to select a digit displacement in the CRT which corresponds to multiplying the scanned spectrum by a given power of 2. Any desired multiplier may be composed of a few such contributions (e.g. to 1 % accuracy in four properly selected positive or negative steps). Thus it takes a few seconds to add into the memory a complete multichannel spectrum multiplied by an arbitrary positive or negative factor.

An unknown pulse distribution received from a detector (switch in positionM) may be analyzed

immediately into a set of standard spectra stored on film, by subtracting each standard distribution from the measured spectrum a sufficient number of times to clear the memory. A stop-signal, produced when a particular channel is cleared, may be used for automatically adjusting coefficients and programming the film transport; thus the unknown spectrum could be decomposed in about one minute into several hundred calibration spectra. The multipliers are totalized and displayed on lamp indicators.

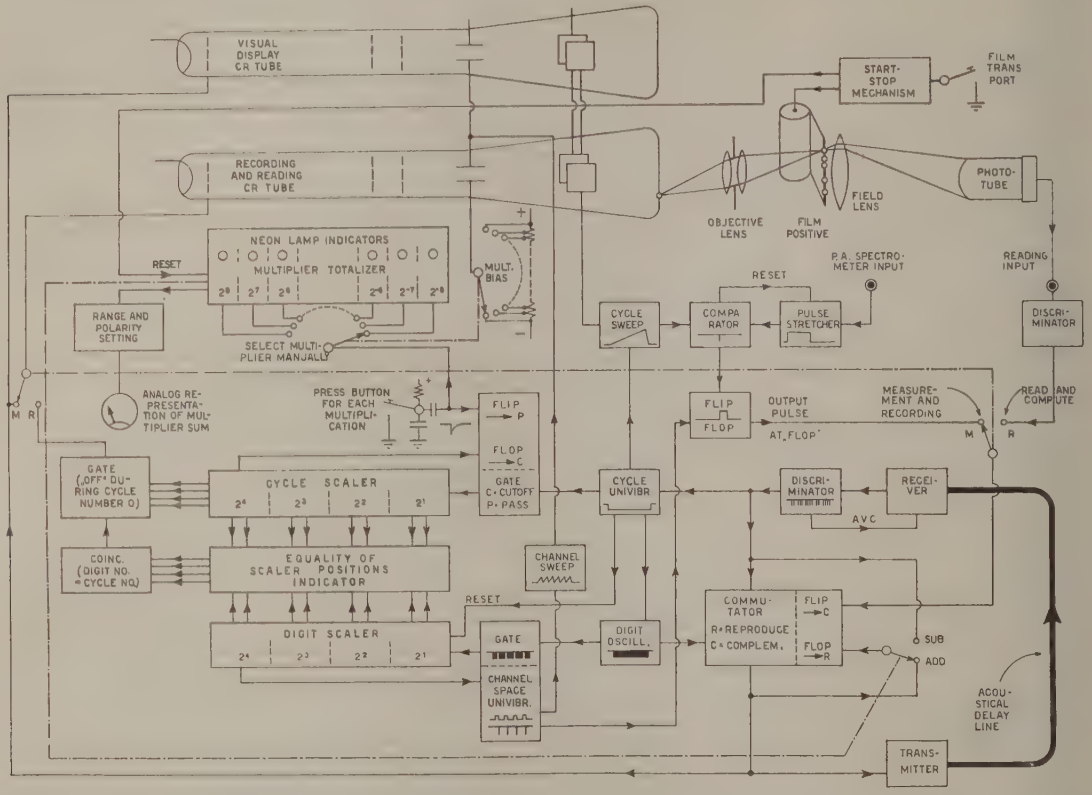

Fig. 4. Proposed photoelectric method of RE-reading multichannel information, stored on film, back into the memory of the same spectrometer which was used in the original measurement, with provision for semiautomatic computing of linear combinations. Note: A pulse at an input denoted "Flop" in this diagram switches the circuit in the respective box into the quiescent state, provided that a "Flip" pulse had previously thrown it into the opposite state.

The proposed scheme is indicative of the relatively few additional components required to convert a storage-type spectrometer into a computer capable of reducing the data.

DISCUSSION

Åström pointed out that in programming a computer it was necessary to make some preliminary assumptions about the nature of the result. In a self-contained, purely automatic system, significant deviations from the expected process might well be overlooked if provision were not made for human intervention at various stages of the experiment. Considerable scientific advancement had often arisen from human assessment of the unexpected.

Maeder agreed that it was valuable to present records in a form readable by the observer but considered it even more advantageous if they were at the same time suitable for operating a mechanical brain.

Kowarski remarked that in the arrangements described by Maeder, the detecting device fed directly a number of automatic data processing stages the output of which was presented to the observer at the end of the chain. This sequence was perhaps proper to branches other than high energy physics. In the latter the observer should intervene after the detector stage in order to assess the data, exercise selection and determine the nature of following computations after which he would be given the final result. This technique of intervention of the observer was used in the photographic evaluation instrument under construction at CERN.

Quercia asked if complicated devices of the type discussed were less likely to be necessary in the field of high energy physics where the spectra encountered were simpler than those met in low energy spectroscopy. In particular were not pulse height analyzers with say 20 channels and short dead time preferable to expensive, complicated instruments of 100 or more channels.

Goldschmidt-Clermont replying to Quercia agreed that high energy spectra were simpler but indicated the existence of a definite need for instruments recording their output in digital form.

For instance, if the pulses of each counter of a meson telescope were recorded in digital form, a computer could be used to investigate their mutual relationships in height and time. The information recorded in a single experimental run could thus replace the otherwise lengthy adjustments of the discrimination levels of the individual counter.

Hutchinson pointed out that in the high energy field, not only had pulse heights from various detectors to be considered but many observables which could be codified into pulse heights and then applied to standard analysers or tapes. Instanced were magnetically analysed momentum or time of flight measurements. In dealing with rare events of rather unknown character it might be necessary to scan simultaneously a range of variables from different types of detector and in order to correlate them the conversion technique could be very useful.

Von Dardel referring to the optimum number of channels in a pulse height analyzer for use in high energy work observed that a flexible apparatus was desirable. A nominal 100-channel device that could be operated as two independent 50-channel or ten 10-channel analysers would be very useful.

Maeder indicated the flexibility of 100-track film in this respect.

Kandiah stated that little modification would be required to convert a currently available type of 100-channel analyser into a ten 10-channel device or five 20-channel etc., but pointed out that the sorting of a pulse into a particular address would mean a closure time of perhaps 40 μs to all other addresses. Requirement of real random and independent access to different sections of the store in an analyser would call for very complicated changes in circuitry. Appendage of intermediate stores between the input and the main memory did not provide the correct solution but merely served to regularize the random occurrence of input pulses, for presentation to the memory.

Hutchinson suggested the use of large numbers of single-channel analysers in conjunction with photographic or magnetic tape recording instead of attempting the complexity of subdivision of a multichannel analyser. This arrangement would be applicable to experiments where a number of variables were all required in coincidence in a region of values for subsequent analysis and also to experiments where a particular value of several variables was needed to trigger promptly other parts of the apparatus.

Merrison pointed out that the choice of a pulse height analyser for use in high energy work was not determined primarily by the size of its number of channels but by its reliability in operation.

NUCLEAR INSTRUMENTS 2 (1958) 136–138; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION III.2

SPEED OF ANALOGUE ELECTROSTATIC STORAGE

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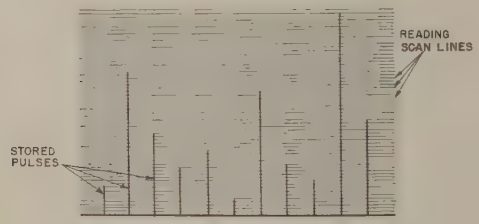
The pulse height analyzer for which the device to be described was developed is based on the principle of storing incoming pulses in a CRT in order to try achieve better resolving time.

Fig. 1 shows the storage pattern. Time increases to the right. Each time a pulse occurs the horizontal deflection is halted so that the



METHOD OF READING EMPLOYED IN PULSE-HEIGHT ANALYZER STORAGE TUBE

Fig. 1

pulse is recorded as a vertical line. After each pulse the beam is stepped to the right by a unit amount so that although pulses appear randomly in time they are displayed equally spaced on the CRT.

The horizontal lines represent the manner of reading out the stored pulse pattern. The same beam that wrote the pulses is subsequently swept horizontally across the stored pulses. Each time it crosses a stored pulse a signal is generated on the pickup plate in front of the tube. The number of these signals generated in each horizontal scan is counted by a scaler and the count transferred to the digital memory system, being added to the previous total therein. Each horizontal line is scanned in a similar fashion.

Thus the memory stores the integral number of counts in each channel and the differential count per channel is obtained by subtracting the total in each channel from the preceding one. This system has been developed to 64 channels.

Fig. 2 shows the experimental results and is the pattern obtained on the face of a display tube using for input the signals generated on the pickup plate of the storage tube. The test signal was a damped oscillatory waveform of 100 kc/s so that each complete vertical portion occupies 5 μs. Since the waveform is sinusoidal the beam is moving most rapidly at the centre points of each vertical section and the speed there is calculated to be 1 in./μs. The theoretical limitations of this type of storage will be shown using this figure.

It will be observed that very strong signals are obtained from the storage tube and also

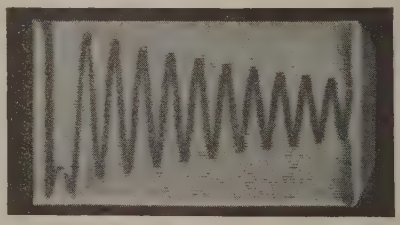


Fig. 2

noticeable is the quantizing action of the tube—the peak of each simulated pulse falls in one channel or the next; with practically no spreading over several channels.

According to accepted theory the CRT beam charges a bombarded spot positively as a result of the fact that secondary electrons are emitted and initially these exceed in number the primaries. In a relatively short time the bombarded spot will reach an equilibrium potential

where the primary and secondary currents are equal. Some of the secondaries will distribute themselves in neighbouring locations so that the positively charged spot will be surrounded by a ring of negative charge.

Of particular interest, for the storage pulse height analyser, is the maximum possible writing speed with a moving beam and in the case of a stationary beam, the time required to charge up a spot on the target. A relationship between these two cases will now be derived, starting from consideration of the simpler case of the stationary beam.

If the diameter of the beam is d and the capacitance per unit area of the target is c the spot capacity will be

$$C = \frac{\pi d^2 c}{4} \quad . \tag{1}$$

The potential V to which the spot charges will be equal to Q/C, where Q is the charge on the spot. This charge is equal to $i\Delta t$, the product of the equivalent charging current and charging time. i is the average net current to the spot and is the difference between the secondary and primary electron currents. For simplicity an average secondary emission ratio of 2 will be assumed so that i is equal to the beam current of the tube.

Hence

$$V = \frac{4i\Delta t}{\pi d^2 c} \quad . \tag{2}$$

The case of a moving beam will now be considered. In time dt the beam traverses a distance ds so that an area $d \times ds$ is charged. In this case therefore, the potential V becomes $i dt/dc ds$ or if $v = ds/dt$ is the velocity of the beam

$$V = \frac{i}{d\,c\,v} \quad . \tag{3}$$

By equating (2) and (3) we obtain the charging time for a stationary beam equivalent in effect to the same beam moving with velocity v.

$$t = \frac{\pi d}{4v} \quad . \tag{4}$$

Referring to the experimental results above, v can be taken as 10^6 in/s. While this is the speed from the photograph, it is possible perhaps to obtain higher writing speeds.

The application of these two cases is as follows: Use of the moving beam corresponds to an integral analyzer of the type discussed in which the complete pulses are stored as vertical lines and scanned thereafter horizontally. Use of a stationary beam corresponds to a differential analyser in which only the peaks of the pulses are stored as dots (again read out by horizontal scanning) with the beam momentarily stationary.

With a moving beam $v = 10^6$ in./s and if a CRT is used that allows a maximum pulse height of 3 inches a resolving time of 3 μs is possible, this being the time to write the full pulse.

Assuming a diameter of 0.009″ for the beam, the charging time for an equivalent stationary beam would be 7 ns, from (4). On the basis of this figure it is possible to calculate the magnitude of the pickup signal and this has been done with good agreement with experimental result.

The question arises as to the importance of very fast resolving time in pulse height analyzers. With pulsed accelerators we are at present limited to analyzing one pulse per burst using analyzers employing pulse height to time conversion. If it becomes necessary to analyze more than one pulse per burst this storage method appears to offer considerable promise.

Several inherent problems require investigation. In scanning across a pattern of dots it is possible to miss a pulse completely or count it twice; if this problem can be surmounted it would appear feasible to build an analyzer with a resolving time of 7 ns. Until this is done the electrostatic storage method appears to have a limit of about 1 μs.

DISCUSSION

REAVES stated in reply to a question that as the same beam was used for reading and writing the deflection sensitivity was not an important factor. The device was in reality a voltage comparator; pulse heights were compared with d.c. or slowly changing voltage.

HOOGENBOOM remarked that it was possible to avoid double counting of the dot by delaying the signal received from a given line and taking it in anticoincidence with the signal from the next line.

REAVES acknowledging that originally the readout had been attempted in a vertical fashion to avoid double pulsing stated that one drawback of this method was a poor signal to noise ratio. Various solutions had been tried such as modulating the scan with a high radiofrequency. The horizontal scan was better.

NUCLEAR INSTRUMENTS 2 (1958) 139–145; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION III.3

FAST LOGICAL BINARY ADDING NETWORK

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1. Introduction

The purpose of this paper is to describe a logical binary adding network that illustrates a new method of coordinating arithmetic operations with memory reading and writing operations in an electronic computer. The adding network to be described was one that was developed at the National Bureau of Standards in connection with an experimental pulse-height analyzer system[1,2]). The technique illustrated by this adding network may also be useful in other information handling systems.

Conventionally, when an arithmetic operation is to be accomplished in an electronic computer the complete number or word to be operated upon is first transferred from the memory into the arithmetic unit. Then the logical operation is performed and the result is transferred as an entity back into the memory. Thus, there are two transfer operations plus the logical operation, or a total of three separate steps that are required to accomplish the desired result. In a system that requires approximately an equal amount of time for each of these steps, the total time required will be about three times greater than the time required simply to read the memory.

In the system to be described, new information is put back into the memory practically instantaneously as the old information is read out. Instead of waiting until the entire number or word is read out of the memory before beginning the arithmetic operation, this operation is done bit by bit in synchronism with the read-out process and the new information is put back immediately, proceeding from the least to the most significant digit. Thus, as each binary digit is read out, the arithmetic logic for all the preceeding (less significant) digits will already have been performed. In contrast to the conventional method, this method requires no longer for the entire operation than is normally required simply to read out information from the memory. The principle of operation is illustrated in the adding network to be described.

2. Block Diagram

Fig. 1 shows a block diagram of the logical binary adding network, together with the memory and control units. A serial-access Williams-type electrostatic memory is shown, and the explanation to be given will apply specifically to this memory. However, another type of serial memory, e.g. an acoustical or electrical delay line might be employed, depending on the particular application.

Synchronous control of the memory and the logical circuits is maintained by a master oscillator, and the frequency of this oscillator determines the rate at which the operation proceeds. One cycle is allotted to each binary digit or "bit". In the experimental system that was constructed the frequency was chosen to be 300 kc/s, giving a time of 3.3 microseconds per bit. The sine-wave output of the oscillator is used to generate three trains of clocking pulses, which have the same repetition frequency but differ in phase. The control scaler shown in fig. 1 counts the first-phase (CP1) pulses and

[1]) H. W. Koch and R. W. Johnston, Multichannel Pulse Height Analyzers, National Academy of Sciences—National Research Council Publication No. 467, pp. 118–130.

[2]) J. H. Reaves, Pulse Height Analyzer System, U.S. Patent No. 2, 804, 606.

provides binary output signals that actuate a number of control gates with various functions. One function of these gates is to select the proper digit from the Addend Register, and in the case when this digit is a "1", an affirmative ADD-ONE signal is supplied to the logical adding network. Another function of the control gates is to provide the memory address information, which in the case of the electrostatic memory consists of signals that control the sweep voltages for the cathode-ray memory tube.

In the multi-channel pulse height analyzer system in which this adding network was employed, each horizontal-scan line in the cathode-ray digital memory tube corresponded to one channel, and there were 64 such lines. Twenty-three binary digits were stored along each line, giving a count storage capacity of 2^{23} per channel. The Addend Register was sequentially provided with numbers to be added respectively to the various channels. Using the logical circuits to be described, each addition was accomplished in approximately 80 microseconds.

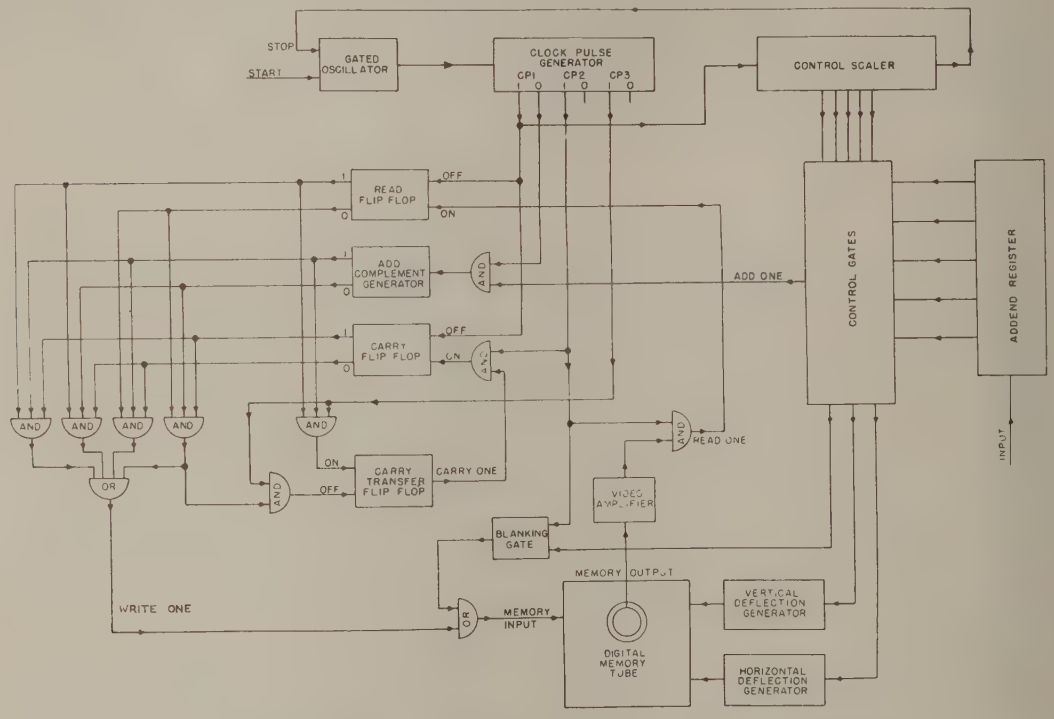

Fig. 1. Block Diagram of Logical Binary Adding Network and Digital Memory.

3. Description of the Operation of the Logical Binary Adding Network

The logical binary adding network comprises the following units shown in fig. 1: Read Flip-Flop, Add Complement Generator, Carry Flip-Flop, Carry-Transfer Flip-Flop and the AND and OR gates that are located adjacent to these units in the block diagram. The details of the operation of this network will be described in this paper, while supplementary information of the other units shown in fig. 1 may be obtained from references[1-3]).

The time allotted to each binary digit (3.3 microseconds) is divided into three parts, t_1, t_2 and t_3, corresponding to the respective durations of three phases of non-overlapping

[3]) F. C. Williams and T. Kilburn, Proc. Inst. Elec. Eng., Part III 96 (1949) 81.

clock pulses. During time interval t_1 the following actions take place:

(1) The Read Flip-Flop and Carry Flip-Flop are shut off.

(2) The Add Complement Generator is held in the "off" position.

(3) The Control Gate unit, which is supplied with information from the Control Scaler, selects the proper output lead from the Addend Register and when a "1" is present on this lead, an affirmative ADD-ONE signal is supplied to the Add Complement Generator.

(4) The Control Gate unit also causes a particular memory location or address to be selected for reading.

The logical network is now in a state of readiness to receive the binary digit from the memory. During time interval t_2 the following actions take place, or may take place depending on the conditions to be stated.

(1) The particular address or digit that has been selected in the memory is read. If this digit happens to be "1" the Read Flip-Flop is turned on.

(2) If a "1" is present on the ADD-ONE lead, the Add Complement Generator is turned to the "on" position.

(3) When a carry signal is present as a result of previous operations, the Carry Flip-Flop is turned on.

The logical AND and OR gates, which are connected in such a manner as to put into effect the rules of binary addition, function as follows:

(a) An affirmative signal is supplied on the WRITE-ONE lead whenever *only one* or *all three* of three input signals are present. The three input signals referred to here are the respective "1" output signals from the Read Flip-Flop, Add Complement Generator and Carry Flip-Flop.

(b) The AND gate that is connected to the ON input of the Carry-Transfer Flip-Flop will turn this flip-flop on whenever a "1" is present on the ADD-ONE lead and simultaneously a "1" is read from the memory. However, due to the clock pulse (CP3) supplied to a third input to this AND gate, this flip-flop will not be turned on until the beginning of time interval t_3.

(c) Similarly, the AND gate that is connected to the OFF input of the Carry-Transfer Flip-Flop will turn this flip-flop off at the beginning of time interval t_3 if a "0" exists on the ADD-ONE lead or a "0" has been read from the memory.

During time interval t_3 a WRITE-ONE pulse causes a "1" to be written into the memory, or if this pulse is missing, a "0" is written. This new digit is stored in the same address that was read only a fraction of microsecond earlier. During time interval t_3 and the subsequent time interval t_1', the carry information is held by the Carry-Transfer Flip-Flop. This information is then transferred to the Carry Flip-Flop at the beginning of the next t_2 time interval, which is designated t_2' and corresponds to a subsequent binary digit.

Upon the completion of time interval t_3, the entire cycle is repeated for the next binary digit. Thus, in as many cycles as there are binary digits in the number taken from the memory, the sum of this number and the number from the Addend Register will be rewritten into the memory.

Typical waveforms for the adding network, and the three-phase clocking pulses are shown in fig. 2. The example that has been selected for illustration uses an augend ending with the decimal digit 9, which appears in binary form (1001) as the signal labeled "Read-One Pulses from Memory". (It should be noted in this instance that the absence of a pulse corresponds to a zero.) The addend in this illustration is 3, which appears in binary form (11) as the Add Flip-Flop output signal. The sum 12 appears in binary form (1100) as the memory input signal. In comprehending fig. 2 one must remember that the least significant digit is on the left since it occurs first; the graph is plotted in the customary manner with time increasing to the right.

In the example illustrated in fig. 2 both the Read Flip-Flop and the Add Flip-Flop are turned on during the first t_2 time interval,

which corresponds to the least significant binary digit. For this combination the arithmetic logical circuits accordingly supply a "0" various parts of this diagram will be explained and identified with their counterparts in the previously described block diagram, fig. 1.

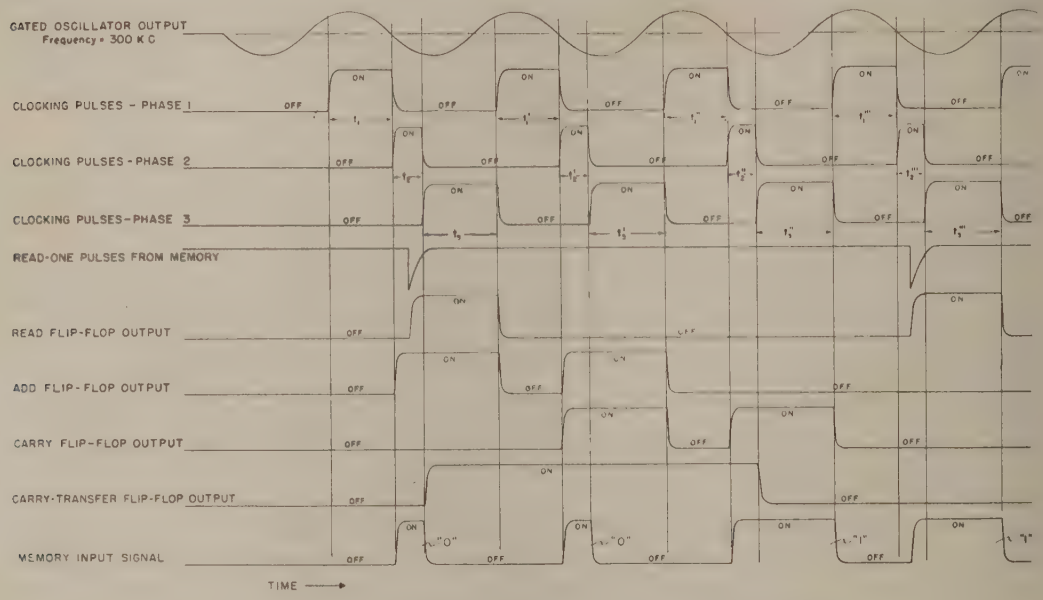

Fig. 2. Control and Signal Waveforms in Logical Binary Adding Network.

write signal to the memory, and the Carry-Transfer Flip-Flop is turned on. The next first-phase clock pulse (CP1) turns off the Add Flip-Flop and the Read Flip-Flop, but the Carry-Transfer Flip-Flop remains in the "on" condition. During time interval t_2', which corresponds to the next binary digit, the Add Flip-Flop and the Carry Flip-Flop are turned on, and for this combination a "0" is again supplied to the memory. Time interval t_2'' finds only the Carry Flip-Flop in the "on" condition, and therefore, a "1" is supplied to the memory and the Carry-Transfer Flip-Flop is turned off by the subsequent third-phase clock pulse. Finally, during time interval t_2''' only the Read Flip-Flop is turned on and consequently a "1" is written back into the memory.

4. Circuits employed in the Logical Binary Adding Network

The complete circuit diagram for the logical binary adding network is given in fig. 3. The In relating this circuitry to the previous description of the operation based on the block diagram, zero potential on a "1" output lead corresponds to the "on" condition whereas a potential of minus 8 V on this lead corresponds to the "off" condition. The reverse applies for a "0" output lead.

The Read Flip-Flop includes tubes T_1, T_2, and T_3. Normally the Read-One input is held at ground whereas the CP1(1) input is normally at approximately minus 8V. The End-Reset Input, whose only purpose is to turn this flip-flop off after the completion of an addition, is likwise normally at a potential of minus 8 volts. Thus, normally the left-hand side of T_1 is conducting, and in this condition the flip-flop is said to be in the "off" state. A negative pulse at the Read-One input is applied to the grid of T_1 through diode D_1 and turns the flip-flop on. The mechanism by which the flip-flop is turned on is as follows: when the left-hand side of T_1 is cut off its cathode voltage drops, thereby also cutting

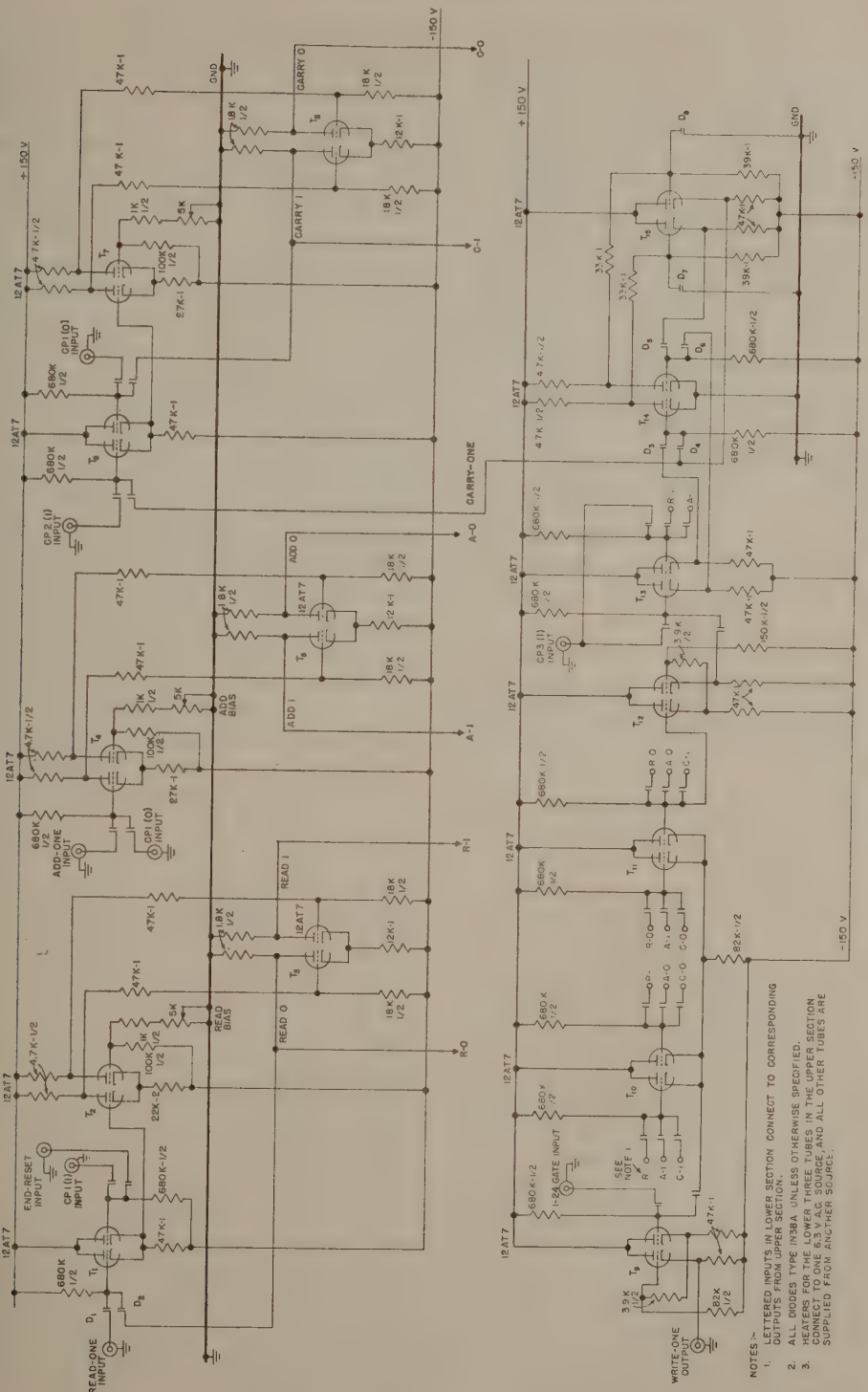

Fig. 3. Circuit Diagram of Logical Binary Adding Network.

off the left-hand side of the difference amplifier T_2 and also the right-hand side of the difference amplifier T_3. Conduction in T_3 is switched to the left-hand section and the resulting drop in the plate voltage of this section causes a negative voltage to be applied to the grid of the left-hand section of T_1 through diode D_2. This diode, which is connected with diode D_1 to form an OR gate for negative signals, acts to hold the flip-flop in the "on" condition. In this "on" condition the R-1 output lead is at ground potential whereas the R-0 lead is at approximately minus 8 volts. The Read Flip-Flop remains in the "on" condition even after the input pulse has expired, but is turned off by the next first-phase clock pulse, which is applied at the CP1(1) input.

The Carry Flip-Flop consists of tubes T_6, T_7, and T_8, and is similar to the Read Flip-Flop except that it is designed to be turned on by a positive-going pulse at the CP2(1) input (provided an affirmative CARRY-ONE signal is present) and turned off by a negative-going pulse at the CP1(0) input. Although for clarity the block diagram shows the Carry Flip-Flop as being turned off by a positive-going CP1(1) pulse, this result was accomplished more conveniently in practice by using the negative-going CP1(0) pulse.

The Add Complement Generator comprises tubes T_4 and T_5. There is no need to use a flip-flop circuit in this instance since the proper output pulse waveform can be obtained more simply using an AND gate that combines the ADD-ONE signal and the CP1(0) pulses. The purpose of this gating is to remove very short pulses that appear in the ADD-ONE signal during time interval t_1 because of the switching that takes place in the control gates.

The four AND gates used in deriving the WRITE-ONE signal are the ones shown in the grid circuits of the dual triodes T_{10} and T_{11}. The OR gate associated with these AND gates consists of a common connection of the four cathodes. A positive-going output at the cathodes results if the proper combination of signals is applied to any one of the four AND gates.

The purpose of the AND gate shown in the grid circuit of the right-hand section of T_9 is to make impossible the generation of a WRITE-ONE signal during sweep fly-back in the memory tube, as the resulting spraying of electrons across the screen during fly-back destroys the information in the memory. The purpose of the left-hand section of T_9, which is connected as a cathode follower, is simply to provide a low impedance output for the WRITE-ONE signal.

The Carry-Transfer Flip-Flop consists of tubes T_{14} and T_{15}. The particular circuit used here is of a different type from that used for the Read and Carry Flip-Flops; however, in principle either type could have been employed for any of these three applications. The manner in which this flip-flop functions is as follows: a positive-going pulse applied to the left-hand grid of T_{14} through diode D_3 causes this section to conduct, and the resulting drop in its plate potential is transmitted via the voltage divider, cathode follower T_{15}, and diode D_5 back to the right-hand grid of T_{14}. Similarly, this negative-going voltage applied to the right-hand grid of T_{14} causes a positive-going voltage to be applied to the left-hand grid of this same tube through the "holding" diode D_4. This diode and diode D_3 are connected to form an OR gate for positive signals. The Carry-Transfer Flip-Flop is similarly turned off by the application of a positive-going pulse to the right-hand grid of T_{14} through diode D_6, and is held in this "off" condition by diode D_5. The purpose of diodes D_7 and D_8 is simply to clamp the grids of T_{15} at ground potential.

The Carry-Transfer Flip-Flop is turned on if at the beginning of time interval t_3 both the Read Flip-Flop and the Add Flip-Flop are in the "on" condition. This required coincidence is detected by the AND gate shown in the grid circuit of the right-hand section of T_{13}. Once the Carry-Transfer Flip-Flop is turned on, it will stay on until both the Read Flip-Flop and the Add Flip-Flop are found in the "off" condition during a subsequent time interval t_3. The latter coincidence is detected by the AND gate shown in the grid circuit of the left-hand section of

T_{13}, which has one of its inputs supplied through cathode follower T_{12} from the AND gate at the right side of T_{11}. In practice this result could have been accomplished more simply using an arrangement similar to that shown with the right-hand section of T_{13}, but the circuit that is shown follows the block diagram, which requires one less AND gate input.

5. Conclusions

The adding network that has been described has worked quite well with a bit time of 3.3 microseconds. Since the design employs direct coupling throughout rather than pulse type circuits, and uses a flip-flop for carry-signal delay instead of a delay line, the speed of operation may be varied quite simply by changing the frequency of the master oscillator. The maximum speed will be limited either by the electrostatic memory or the transient response of the circuits that have been shown, or by both. With only minor modifications a speed of two microseconds per bit appears achievable.

With respect to the electrostatic memory actually employed with this adding network, some difficulty was experienced in eliminating errors resulting from noise carried into the system by way of the power lines. By extensive filtering and shielding of all circuits connecting to the memory tube, especially the video amplifier, the number of errors was greatly reduced. With further development the trouble conceivably can be completely eliminated.

The operational principles illustrated by the logical adding network that has been described might possibly be applied to serial type memory systems operating at a much higher speed than two microseconds per bit when such memories are developed. Also, the principle of carrying out logical operations bit by bit nearly simultaneously and in synchronism with memory reading and writing operations might also be advantageously employed with parallel access memory systems.

Discussion

In reply to a question, Reaves stated that the storage of digits was similar to that described by Williams and Kilburn viz. a dot for a zero and a dash (or elongated dot) for a one.

NUCLEAR INSTRUMENTS 2 (1958) 146–153; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION III.4

INSTRUMENTS FOR THE ANALYSIS OF TRACK CHAMBER PHOTOGRAPHS BY DIGITAL COMPUTER

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Presented by Y. GOLDSCHMIDT- CLERMONT

1. Definition of the Purpose

The track instruments such as the cloud or the bubble chamber, particularly when used in a magnetic field, give detailed and relatively complete information on the high-energy events they are made to record. When used in the beams of high-energy accelerators, they accumulate rapidly large numbers of photographs of interesting events. Many measurements and computations are needed to extract significant data from each photograph. Yet the information required on the high energy events is usually statistical in nature, that is obtainable only from the detailed analysis of a large number of events. The rather slow methods which, up to now, have been in use, for instance in cosmic-ray physics, cannot deal with the present requirements.

If, in the succession of the operations used in these "classical" methods (event recognition, measurements performed on the selected tracks, spatial reconstruction, calculation of geometrical corrections, calculation of physical magnitudes), we try to locate the slowest bottlenecks, we notice that these do not reside in the visual scanning, nor in the operation of bringing a cross-wire in coincidence with a visually selected point of the track. The really laborious part begins with the reading of the coordinates thus obtained, and the subsequent data processing leading to such results as the numerical values of curvature, true coordinates in space etc.

It is precisely in this domain that digital computers may prove most helpful, thus giving us a reasonable hope of catching up with to-day's speed of production of significant photographic data. To that effect, the information contained in the photographs will have to be translated into the computer code, and recorded on punched cards or tape, or on magnetic tape. The translation will be performed by an observer using specially designed measuring instruments. Several proposals, mostly unpublished, for such instruments have recently been put forward in various high-energy laboratories; in this paper we discuss the instruments at present under development in CERN.

2. Plan of this paper

To begin with, we define the problem by giving a brief description of the chamber photographs together with some considerations concerning their accuracy, corrections, etc... (input data). As bubble chambers seem to be particularly well adapted to the problems of high-energy physics now being studied with the accelerators, some of the numerical values quoted will refer to them. We assume that the reader is sufficiently familiar with high-energy physics so as to dispense with the description of the desired output data such as centre-of-mass angles, Q values, masses, etc.; in this connection we point out the necessity of using several measuring instruments of various degrees of accuracy.

We follow up with a description of the "Logic", that is the succession of interlinked operations which are needed in order to go from the films to the final results.

The actual design of the instruments seems to depend on a rather small number of basic decisions, some of which are examined in the last sections of the paper where the CERN instrumentation programme is described. For instance, the designer must decide how to divide the operations between the instrument, the observer, and the computer, with respect to the special abilities of each. As the role of the observer is progressively reduced, his attention should be called upon only where his intervention is essential, and all accessory operations such as film centering should be reduced to the minimum. At the end of the paper we mention some of the stages of this evolution.

3. Tracks of Nuclear Particles

In a typical fruitful chamber expansion there is an "event" to be studied, consisting usually of a number of "tracks", sometimes converging to an "apex". The tracks are defined by a string of bubbles (bubble chamber) or droplets (cloud and diffusion chamber) of somewhat unequal sizes, and spaced at random.

If the chamber is used without a magnetic field, the tracks of fast particles are approximately straight. Slower particles show multiple scattering. In a homogeneous magnetic field, the tracks have helicoidal shapes modified by multiple scattering. Slow particles losing an appreciable fraction of their energy in the chamber show multiple scattering about a helix of decreasing radius of curvature.

If the thermodynamical parameters of the chamber are kept sufficiently constant, the bubble or drop density along a track gives a measurement of the particle velocity. Methods for the measurement of track density and multiple scattering are not examined in the present report.

4. Accuracy: Location of Bubbles

The precision of the instrument will be evaluated with reference to the bubble chamber, taken as an example. In the present state of the art, turbulence, thermal distortion and other faults affect the accuracy; we assume, however, that technical advances, soon to be achieved will make the bubble chamber as precise as the cloud and diffusion chambers.

The bubble size depends on the operation of the chamber. A typical diameter is 200 μ. The centre of the bubble can be located with a further definition of, say, 1/10, so that its position in space can be defined with an accuracy (absolute) of about 20 μ. For an overall size of the chamber of the order of 20 cm, the relative accuracy is about 1 part in 10^4. The same or a slightly better relative accuracy is achieved in cloud and diffusion chambers.

On the inside faces of the chamber windows are "fiducial marks" used as reference points. Their exact position is accurately known.

5. Accuracy: Photography

The chamber is photographed with two or more cameras to provide "stereoscopic views" that will allow spatial reconstruction. In some exceptional cases, the angle between the optical axis of two cameras is of 90°, making the reconstruction especially easy. The angle is usually much smaller, and the films are often parallel to the front glass of the chamber, so that the two views are similar in appearance: roughly, each view shows an approximate cylindrical projection of the events on the front glass, the differences between the views (parallax) are used to measure distances from the front glass, or "depth".

The optical quality of the lens and the grain size of the film can be chosen to introduce little loss in image definition. Each bubble will usually be represented now by a spot (diffraction rings). The photography is taken with a reduction of, say, 1/5 to 1/10, so that the bubble size, on the film, is about 20 μ and the position of the bubble centres can be defined with an absolute accuracy of the order of 2 μ.

The photography introduces a number of distortions. Some are systematic, others vary from picture (or "frame") to picture. The systematic distortions are due to the defects of the lens, to the gas expansion in some cloud

chambers, to the refraction of the light in the successive media traversed (chamber liquid, windows, air, etc...). The distortions which may vary from frame to frame are due to shrinkage of the film during development, and to local thermal changes of the refractive index of the liquid in the chamber, or to the effects of convection currents.

The film is not always perfectly lined up in the camera so that the position of the droplets or bubbles must be taken with respect to the fiducial marks.

6. Accuracy: Output Data

We have indicated, in section 4, the maximum accuracy to be expected from bubble chamber measurements. Similar figures apply to the gas chambers. Maximum accuracy is necessary for a large number of experiments, such as measurements of Q values, identifications of decay modes, measurements of ranges.

It is useful to point out that there are experiments where less accuracy is required. For instance, if an angular distribution of emission of a decaying particle is studied, and if the total number of measured events is limited to a few thousands, it is not necessary to measure each angle of decay with a precision much higher than a few percent. When less accuracy is sought, the photography, the measuring instrument and the methods of analysis can be less refined, and the economies made in this direction should enable to gain on the speed of measurement. Bubble chamber pictures accumulate very fast, and slow methods should be reserved for the hard cases.

7. Logic

Fig. 1 represents the chain of operations to be performed to extract the required information from the chamber photographs. The entire programme is divided in steps or sub-programmes shown in boxes. Circles under the boxes indicate the information which is added from outside the chain. The table shows, below, the part of the chain taken over by the digital computer in the systems described in some detail later. We first briefly examine the successive steps of the chain.

7.1. EDITING

In a typical chamber experiment, only a fraction of the pictures show interesting events. An observer scans the successive pictures, and selects those containing significant events. When the events are relatively rare, this operation is best made with a simple scanning device (such as ordinary projectors) and only the selected pictures are transferred to the measuring instrument. This editing must be made with more than one stereoscopic view.

As indicated on the table, the essential role of the observer is the recognition of significant events, usually against a rather high background of random tracks.

It is very challenging to attempt to automatize also this part of the process, thereby doing away almost entirely with the human observer. At least in some categories of experiments it is conceivable that a mechanism could be made to recognize certain types of events.

It is too early to predict how such an instrument would work, but some indications can be found in related fields. There exist already mechanized microscopes counting all tracks of a given direction in a nuclear emulsion[1,2]), and electronic automatontranslating printed letters and numerals into a computer code[3]).

7.2. MEASUREMENT

As a first step towards digitization, we have to measure the coordinates of significant points. The design of the measuring instrument is much influenced by the adjoining links of the chain. The measurement must be coded and recorded according to the input device of the digital computer, on punched cards or tape (other input devices will become available in the future). The punching is done by hand for

[1]) S. J. Goldsack and H. B. van der Raay, J. Scient. Instr. **33** (1956) 135.

[2]) E. Amaldi, C. Castagnoli and C. Franzinetti, Nuovo Cimento, Ser. X Vol. 4 (1956) 1165.

[3]) ERA (An electronic reading automaton) Electronic Engineering **29** (1957) 189.

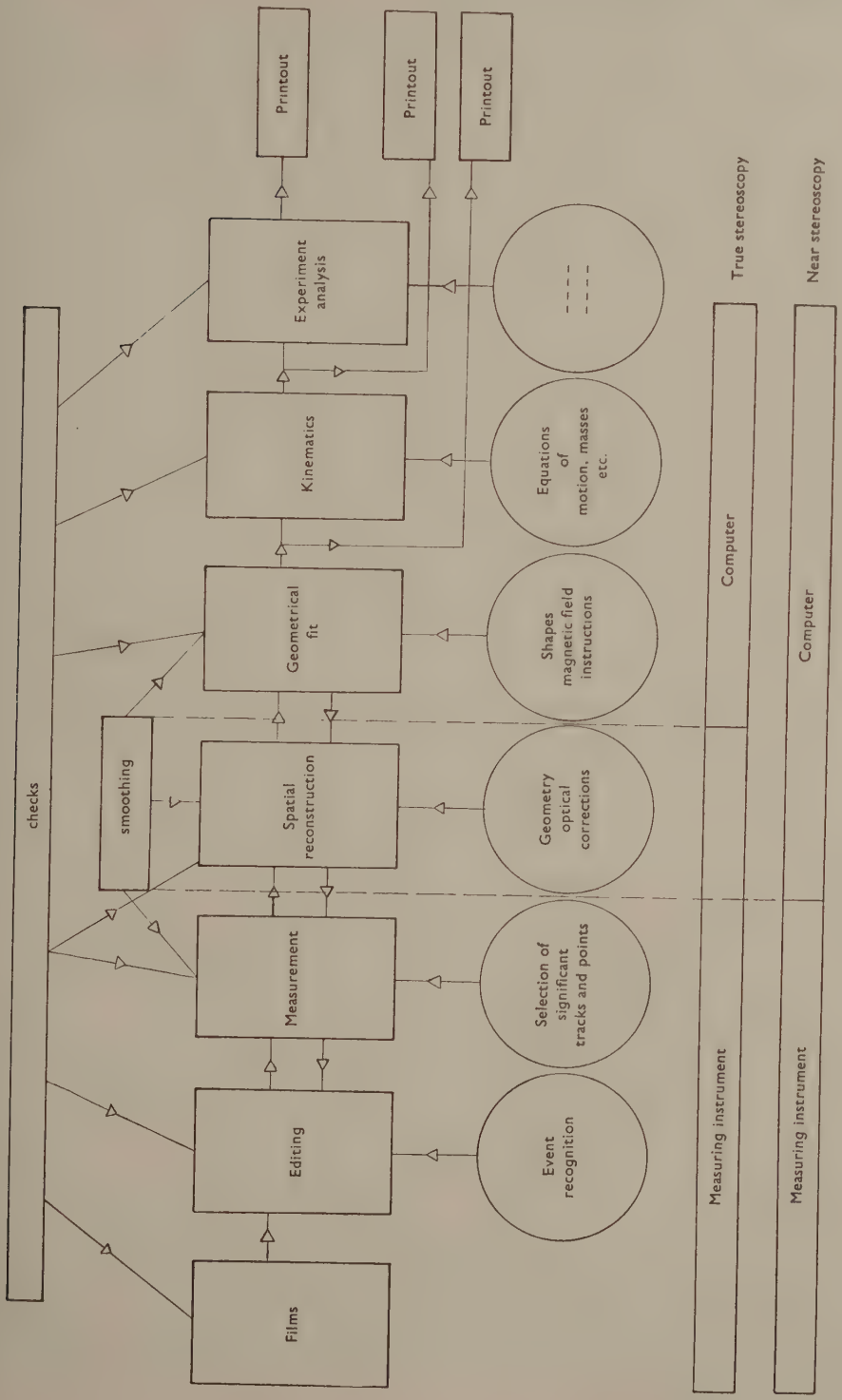

Printout
Printout
Printout
Experiment analysis
Kinematics
Geometrical fit
Spatial reconstruction
Measurement
Editing
Films
checks
smoothing
Event recognition
Selection of significant tracks and points
Geometry optical corrections
Shapes magnetic field instructions
Equations of motion, masses etc.
- - - -
- - - -
Measuring instrument
Computer
True stereoscopy
Measuring instrument
Computer
Near stereoscopy

Fig. 1

some of the measuring instruments (for instance, when the curvature is estimated with a template), but of course, time and accuracy are gained most when the measuring instrument itself has digitized output. As the electronic computing is appreciably faster than the measurement, the punched record turns out to be a convenient intermediary storage, where a large number of results can be accumulated, to be later processed in rapid succession.

7.3. SPATIAL RECONSTRUCTION

Each of the chamber pictures is a conical projection of the spatial event. To reconstruct the spatial event from its projections is a straightforward geometrical problem. The outside data needed are the positions of the cameras with respect to the chamber and, in addition, corrections for the aberrations of the lenses, and for the effect of refraction of the light in the liquid and the window.

Although simple in principle, the geometrical problem is not easy to solve explicitly. Several methods have been used for the Wilson chambers and can be adapted for use with digital computers. But the latter open the possibility of using new methods. It turns out that the design of the measuring instrument depends very largely on the method of spatial reconstruction.

7.4. GEOMETRICAL FIT

To the tracks now reconstructed in space must be assigned a geometrical shape, such as a straight line or a helix, of which we want, at certain points, the direction (tangent) and radius of curvature. Some methods of measurement (templates) already use a geometrical fit, and also some methods of spatial reconstruction. When the operation is done with points reconstructed in space, the computer can use a least-squares fit. Special weight can be given to singular points, such as the apex of an event, where the fit must make the tracks meet. If the magnetic field is not homogeneous, or if the particle loses an appreciable fraction of its energy along the track, appropriate corrections are introduced. Thus, the information to be fed from outside the chain concerns the expected shapes of the tracks, a table of values of the magnetic field, and instructions on the output data needed.

It is useful to print out (punched tape) the data obtained at this point (directions, helix parameters, radia, etc...). They are then available for use with different programmes of analysis: this procedure allows for much flexibility.

7.5. SMOOTHING

A smoothing is introduced when the track images are subjected to a geometrical fit: irregularities, small angle scattering, etc. are eliminated.

As indicated in fig. 1, some smoothing may occur at several places: when the actual measurement is made, the observer may select the chosen point as an average along a segment of the track; in some methods of spatial reconstruction, a geometrical fit is made for each stereoscopic view, etc.

Two aspects of the smoothing may be pointed out:

(1) smoothing eliminates part of the random errors, and the method used sometimes yields an useful estimate of the residual errors (e.g. the least square fit readily yields a value of the residual error);

(2) when a template is fitted to the track images, the smoothing is made by eye. It appears that the eye judges rather accurately and swiftly whether a fit is good enough. This must be kept in mind for the design of the measuring instrument, particularly when speed is important.

7.6. KINEMATICS

From the measured directions of tracks, curvatures, etc., the required dynamical quantities, such as Q values, centre of mass angles, etc. are computed. Here, we introduce in the chain the equations of reaction, kinematics, assignments of masses etc. Often an event is liable to various interpretations, and the

corresponding mass assignments must be used in successive computations. It is useful to perform this operation as an independent computing programme, using as input data the tape printed out after the geometrical fit. The results, of course, are also printed out on punched tape, and tabulated for inspection.

7.7. EXPERIMENT ANALYSIS

Again in a separate computing programme, the results from the kinematic analysis on a number of events are assembled to take weighted means, to draw histograms, to make maximum likelihood estimates etc. As in the preceding step, it may be desired to subject one collection of events to several analyses, and thus to feed it to the computer with different programmes.

7.8. CHECKS

Human and mechanical errors, if undetected, lead to meaningless results. It is essential to use checking procedures everywhere. For instance, the computer programme may include a detection of points lying far away from their neighbours on a track (one wrong digit). The digitizers must be frequently checked by a built in routine. The coordinates of the fiducial marks could be taken before and after a set of measurements, and compared. When making a calculation by hand, steps are often retraced and crosschecks are used. The same should apply to the digitized procedure, and even more so, if the results coming out of the "black box" are to be believed.

8. Solutions Adopted for CERN Instruments: Spatial Reconstruction

In the instruments examined here, no attempt is made to replace the observer in the two essential first steps of analysis, the scanning and selection of significant tracks. The choice of a method to perform the next two operations, the measurement and the spatial reconstruction, is decisive for the design of the instrument.

The measuring instrument may be built to record directly coordinates in 3-dimensional space (True stereoscopy, fig. 1). For instance, it can make use of the human ability to locate objects in 3-dimensional space. Each eye of the observer is given a virtual image of one of the views, as in the classic stereoscope, and also the image of a moving pointer, which is brought manually in 3-dimensional coincidence with successive points on the tracks. Similar instruments are used, for instance, to construct maps from aerial photographs. They have to fit rather strict mechanical and optical requirements: the pointer should be seen in coincidence with a virtual image of the tracks; individual adjustments must be provided for each film holder to achieve the spatial coincidence of the two images. It appears rather difficult to build an automatic device relieving the observer from the tedious task of centering exactly the pointer on successive points of the track. These drawbacks are balanced, to some extent, by advantages over other methods: the observer may get a better understanding of the spatial relationships between various tracks of the event, he may find quick and easy ways to make a coincidence in three dimensions, each event needs only to be measured once.

The computer can relieve the observer from the task of spatial reconstruction, using two-dimensional measurements directly taken on each of the steroscopic views, and the geometrical relationships derived from the relative positions of the chamber and the cameras ("Near stereoscopy", fig. 1). The stereoscopic views of an event are examined in succession. The inconvenience of this repetition is appreciably compensated by the possibility of using an automatic centering device, described below. The role of the observer is thus limited to a coarse driving of the pointer along 2-dimensional views of each track, while a servo-mechanism provides a precise centering. Coordinates are recorded on the fly. No precise alignment of the films is necessary, if the measurement of each view starts by the recording of coordinates of fiducial marks. The sequential measurements of the stereoscopic views make it impossible to associate the coordinates of points corresponding, on each view, to the images of the same spatial

bubble: the structure of the tracks must be ignored. It can be shown that the overall accuracy of measurement can be maintained, provided each event is photographed by three or more cameras, with objectives not on a straight line: the geometrical overdeterminacy compensates the neglect of the track structure. An instrument of this type ("Frankenstein") is in operation in Berkeley[4]). The first CERN instrument, now under development, also uses near-stereoscopy.

All the operations described in this Section are defined in terms of coordinates of points. They can also be based on defining the direction of lines (angles) and their curvature (templates). As the coordinate measurement leads to a higher precision, instruments for this purpose will be described in some detail below.

The human eye judges rather well whether a given line passes through a number of points. This kind of integration is useful for numerous experiments where measurements of ultimate accuracy are not required. A precision protractor or a draughtsman's pantograph can be fitted with rotary digitizers, and used to measure an enlarged projection of the photographs. Such simplified instruments are now studied in CERN, but will not be described further here.

9. CERN Instruments: Digitization

The 2-dimensional coordinates of points are measured by adjusting a pointer on an optically enlarged image of the track. The relative motions of the pointer and the track must be ascertained with a precision of the order of 1 part in 10 000, digitized, and coded. The moving organ can be the film, the objective, or the pointer, and the latter can move at a small or large scale (before or after the optical magnification). The choice again has an important bearing on the instrument design, it is guided by reasons of mechanical and optical accuracy and simplicity, and by ease of adapting an automatic centering device.

The latter requirement practically excludes the motion of the pointer. The detecting organ of the centering device is liable to have an appreciable size and weight, its precise motion should be at the scale of the magnified image, it is better to avoid this mechanical complexity. A choice then remains between the motion of the objective or of the film. The film motion, with fixed objective and detector, has the appreciable advantage of using always the same optical path for the measurement, and therefore of eliminating optical distortions. The motion to be measured is at the scale of the film, and should respect tolerances of ± 2 microns. This is within the scope of precision lead screws and commercial rotary digitizers, or of the linear digitizers described in the next paper. The moving parts, however, must carry the film clamps and transport mechanism, which may be quite bulky.

The CERN instrument† will, instead, move the objective. The film transport and other mechanical accessories are then independent of the precise objective motion. This appreciable mechanical simplification is balanced by some optical problems. The light path no longer remains constant, but extends over a small angular region (with a moving pointer instead, the angular region would be still larger by a factor equal to the magnification). The condenser illuminating the film cannot be of standard design, but should focus light over the whole area covered by the objective in its motion.

10. Proposed Stages of Development

The first instrument will be built in successive steps of increasing automation described in more details in the following paper. In the first stage, now nearing completion, the observer moves the projected image of the event on a screen (about full size). By turning manually the screws driving the cross traverses of the objective stage, he brings successive points of the track in coincidence with a pointer fixed to the screen. A binocular microscope can be used for precise adjustment. With a pedal, orders are given to punch the coordinates. An electric typewriter is used to punch identification numbers and

[4]) Private communication by Dr. L. W. Alvarez and Dr. H. Bradner.

† See fig. 1, p. 155.

coded orders for the computer, the simultaneous typewriting serves as a check.

The next step will change the controls to a steering wheel and velocity pedal, with automatic coordinate registration at regular intervals. The observer will control his machine like an automobile driver. There are reasons to believe that this type of steering will be well adapted to the observer's abilities, and will therefore expedite his task.

In the final version, the observer's strain will be further relieved by an automatic centering device. For instance a television signal can be used to measure the distance from the track to the pointer, and to drive a servomechanism adjusted to zero the signal. The role of the driver will then be limited to a coarse driving along the track and can be expect to be very rapid.

NUCLEAR INSTRUMENTS 2 (1958) 154–163; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION III.5

THE PRESENT STATE OF THE INSTRUMENT FOR EVALUATION OF PHOTOGRAPHS AND FUTURE DEVELOPMENTS

G. VON DARDEL, Y. GOLDSCHMIDT-CLERMONT and F. ISELIN

CERN, *Geneve*

Presented by F. ISELIN

1. Introduction

The aim and background of the instrument for evaluation of photographs, IEP, has been given in the paper by Y. Goldschmidt-Clermont *et al.*†.

A block diagram of the electrical and electronic parts is given in fig. 1. The full drawn lines indicate the extent of the modest first stage of the development program which is now near completion and which will be followed by more ambitious machines indicated by the dotted parts of the diagrams.

In all versions the photograph is projected on to a screen by means of a lens which can be moved on a precision stage to bring successive points of the tracks in the photograph into coincidence with a fiducial mark on the screen. The motion of the mechanical stage in the x and y directions is digitized by use of an optical system developed by L. A. Sayce at NPL[1]) and produced commercially by Ferranti Ltd[2]). In this system, lines of a grating on the movable part pass over a fixed grating to produce moiré fringes detected by phototransistors. Counters are used to record the number of fringes. These counters can add and subtract.

When the cross hair on the screen is in coincidence with a point to be measured the counts of the counter are recorded on punched tape. In addition to the coordinates, numbers giving instructions to the computer can be punched on the tape from a keyboard.

† "Instruments for the analysis of track chamber photographs by digital computer", p. 146.

2. Method of Centering on the Track

In the first model the knobs giving the x and y motion of the projecting lens will be moved manually by the operator. As a first step towards a more accomplished instrument, motors with a wide speed range (about 1 000:1) will drive the knobs (stage 2).

The motion in stages 1 and 2 is achieved by successive displacements in the x and y directions. This will probably be rather inefficient and put undue strain on the operator. One of the important requirements of the IEP is that it be as comfortable as possible so that its efficiency is not limited by the endurance of the operator. It is felt that a good way to achieve this comfort is to imitate the driving of a motor car.

In the proposed car model the operator will bring the crosshair to bear on the desired point of the photograph by control of the curvature of the trajectory with a steering wheel as in a car. How this is done is shown in the block diagram.

The steering wheel controls the speed $d\alpha/dt$ of a motor which turns a sine-cosine potentiometer. This potentiometer provides the control voltages for two variable speed motors which displace the stage in the x and y directions. As the speeds in these directions are proportional to $\sin\alpha$ and $\cos\alpha$ the direction of motion is determined by the position of the sine-cosine potentiometer and since α increases or decreases at a

[1]) L. A. Sayce Endeavour **14** (1953) No. 48.

[2]) J. Guild, The interference systems of crossed diffraction gratings (Clarendon Press, Oxford, 1956); P. J. Farmer, Digital Control (Aircraft Production, July, 1956).

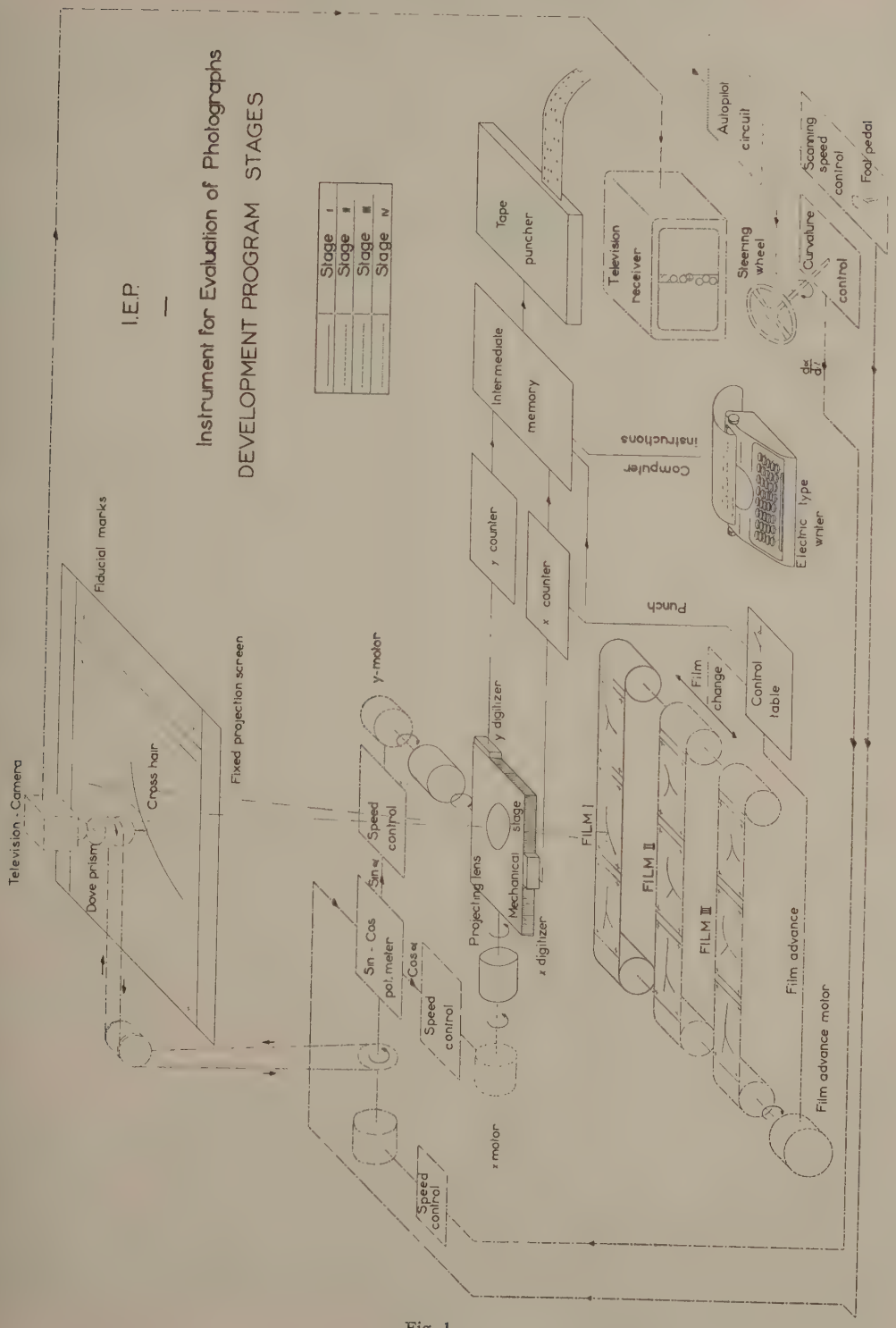

I.E.P.
Instrument for Evaluation of Photographs
DEVELOPMENT PROGRAM STAGES
Stage I
Stage II
Stage III
Stage IV
Television - Camera
Dove prism
Fiducial marks
Cross hair
Fixed projection screen
Sin - Cos pot. meter
Speed control
y-motor
x-motor
y digitizer
x digitizer
Mechanical stage
Projecting lens
FILM I
FILM II
FILM III
Film change
Control table
Film advance
Film advance motor
Punch
Computer
instructions
y counter
x counter
Intermediate memory
Tape puncher
Electric type writer
Television receiver
Autopilot circuit
Steering wheel
Curvature control
Scanning speed control
Foot pedal

Fig. 1

rate determined by the position of the steering wheel, the motion will be along a circle if a constant speed is maintained with the radius of curvature determined by the steering wheel.

The operator will also have at his disposal an accelerator pedal to control the speed of the motion by acting on the voltage across the sine-cosine potentiometer. This voltage is also applied to the potentiometer controlled by the steering wheel so as to avoid a change in curvavature with change in speed.

It is interesting to estimate the expected speed with which the operator can drive along a track in a bubble chamber photograph and still keep to the centre of the track with the desired accuracy. A skilled car driver can easily keep to the centre of the road to within $\frac{1}{2}$ metre at a speed of 100 km/h. By analogy, an accuracy of 1 μ should be feasible for a track speed of 0.1 mm/s on the photograph. If this is borne out in practice there will be no need to stop to center accurately on the track and coordinates could be plotted "on the fly". The operator would therefore drive along the tracks and the instantaneous coordinates would be recorded periodically without action of the operator.

3. Presentation

Proper presentation of the picture is very essential to avoid fatigue. The operator will have in front of him the large scale projection of the picture on the screen with the cross hair, which allows him to move to the right track. For accurate positioning a magnified picture of the region around the cross hair will be necessary. In the first models this is provided for by means of a simple optical system such as a binocular magnifier. Later it is intended to have a television camera pointing at the region around. the cross hair. A television receiver placed beside the operator will thus give a magnified view of the track to be used for accurate scanning whereas the full-scale projection of the picture will serve as a roadmap to localize the proper tracks. A total magnification of 500–1 000 times is to be provided between the photograph and viewing screen of the TV tube.

It is probably convenient to have the picture of the track on the TV always vertical, irrespective of the direction of the track on the projection. Therefore it is intended to insert a Dove prism in front of the TV camera. This prism is linked to the sine-cosine potentiometer by a 2 to 1 reduction link since the picture is rotated 360° for a 180° rotation of the prism.

The last stage of the development program will be a device which will make it unnecessary for the operator to centre the cross hair accurately on the track.

Several systems are possible and require investigation. They would all start from the electrical signal from the TV camera from which information can be derived, in electrical form, on the centre of gravity or median of the track. This information could be used as an error signal to control the motion of the mechanical stage, making a servo system similar to an autopilot, which makes the cross hair follow the track. Another scheme is to abandon centering the cross hair on the track and to digitize the error signal, together with the coordinates of the cross hair, to allow the computer to correct for the error. In yet another scheme the operator would drive along the track in oscillatory fashion and coordinates would be taken at moments when the error signal is zero.

These schemes would reduce demands on the skill of the operator.

At present stage 1 is near completion and only preparations have been made for the subsequent stages of development. A block diagram of the electronics is shown in fig. 2. This part is common to all stages of development.

4. Analog to Digital System

The automatic operation of machine tools has led to the development of several positioning devices. These are digital to analogue systems which convert coordinates into displacements. The present application works in the reverse direction and calls for different electronic circuits from the type associated with machine

tool systems. After consideration of the different possibilities it was decided to use the double grating system developed by L. A. Sayce, and made by Ferranti[2]). This has the major advantage of having a low error of measurement and of being independent of backlash since one grating moves with the measuring table and the other is in a fixed reading head.

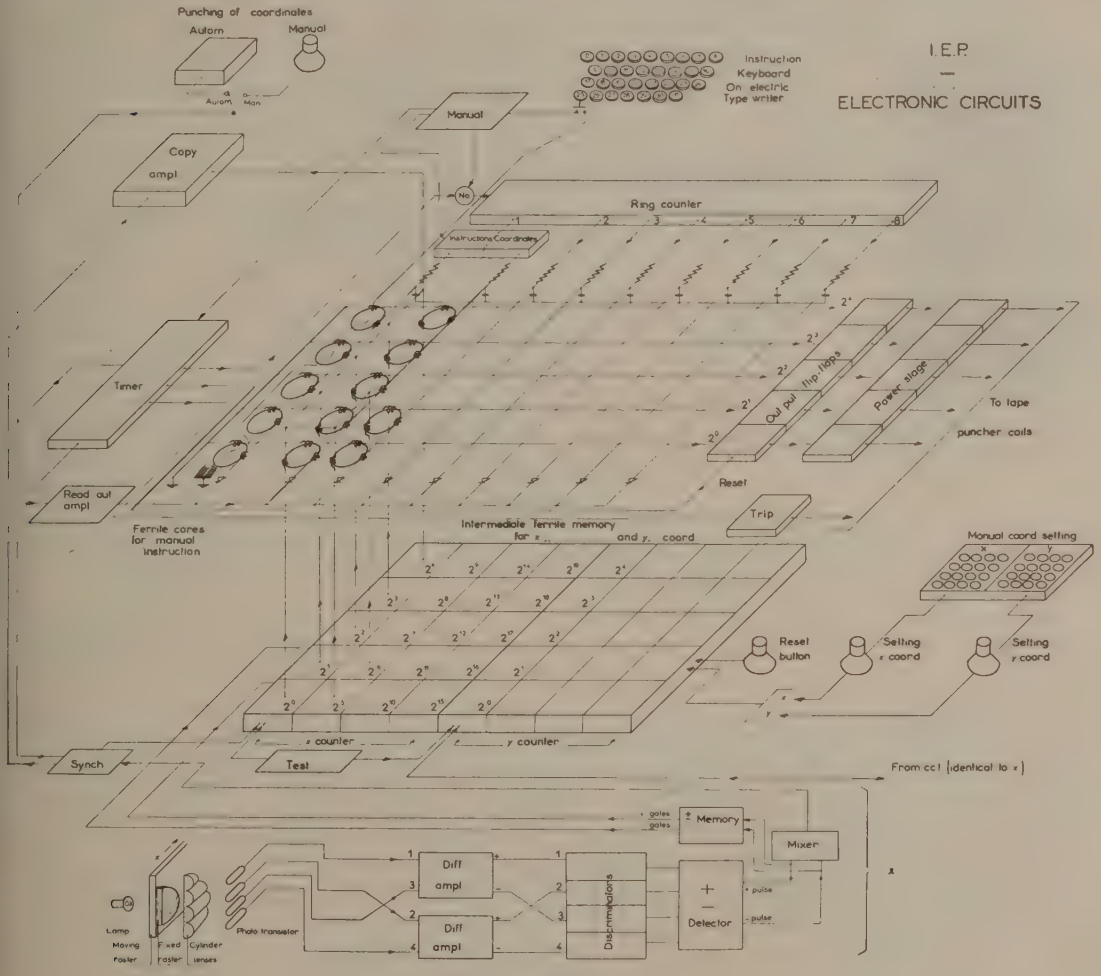


Fig. 2

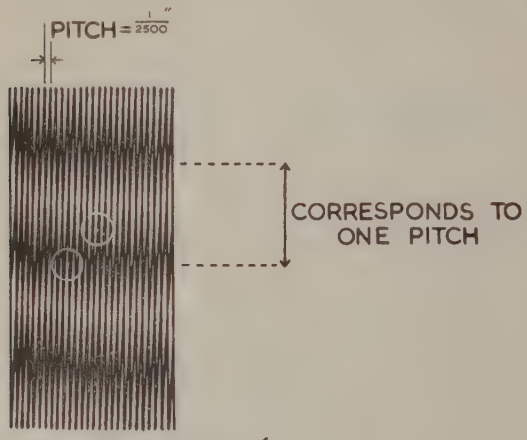
At present the gratings used have a pitch of 1/2 500 inch, (about 10 μ).

The digital information is extracted from the gratings by means of two phototransistors 90° out of phase (circles fig. 3) and at the same time the direction of movement is detected. Four phototransistors are fitted in the Ferranti heads and these are paired to permit compensation for light and temperature effects as well as

4.1. PRINCIPLE

If two ruled gratings are superimposed and one is slightly tilted with respect to the other, there results a moiré fringe pattern in which the light intensity distribution is approximately sinusoidal (fig. 3). The fringe pattern moves in a direction perpendicular to the relative motion of the two gratings.

to extend easily the measuring accuracy to 1/10 000 inch (2.5 μ). A finer grating will soon be available with a pitch of 4 μ with which



PRINCIPLE OF A MOIRÉ GRATING SYSTEM

Fig. 3

measurement to 1 μ could be made. Display of the two signals as a Lissajou figure on a cathode-ray tube would permit extending the accuracy of measuring down to roughly 0.2 μ with the 4 μ grating.

5. Electronics

The electronic circuits have been transistorized in order to reduce considerably the size of the equipment and the heat dissipated. As transistors have a long life, time lost due to failures is kept to a minimum. Plug in cards have been employed, some with printed circuits, to simplify circuit maintenance.

5.1. DETECTORS

The phototransistor signals are taken to two differential amplifiers which eliminate DC components due to temperature or light differences (fig. 4).

A gating circuit determines whether the carriage is moving to the right or to the left (fig. 5). The outputs of the Schmitt trigger discriminator (fig. 2) stages are taken to an *RC* gating circuit so that pulses appear at one output of the sign detector for motion to the left and at the other for motion to the right. These pulses go to the direction memory circuit, which provides the sign control voltages and to the mixer which feeds all pulses (plus or minus) to the binary counter.

5.2. BINARY COUNTERS

There are identical units for the x and y coordinates[3]). Both counters consist of twenty binary flip-flops, each stage being connected to the next through two gates controlled by the sign detector. These gates determine whether the counter counts forwards or backwards. A maximum speed of about 20 kc/s is possible with the simple gate system used which would correspond to a film motion of 2 cm/s with a 1 μ grating. This speed is excessive for accurate scanning of the film but is necessary for rapid motion across the picture. One binary stage, the two gates, the ferrite core (see "intermediate memory") and its driving circuit are mounted on a plug-in card. 40 of these units are provided for x and y. The fig. 6 shows the binary stage with the ferrite and its driving transistor.

5.3. INTERMEDIATE MEMORY

In the more advanced stages 3, 4 it is intended to make measurements "on the fly", without stopping the stage for sufficient time to read out the position of the scaler. An intermediate memory will then be necessary into which the instantaneous position of the counter is copied in a time short compared with the pulse spacing The information from the intermediate memory can then be read out at a rate suitable for operating the tape punch.

This intermediate memory has already been provided on the first model and consists of a matrix of 8×5 ferrite cores each core corresponding to one of the stages of the x and y counters. Each core has three windings, called "copy", "readout", and "output". The copy winding is connected to the binary counter and will carry a current when a short negative voltage pulse—copy pulse—about 20 μs long is applied to the common terminal of the copy windings, if the stage of the counter is in the

[3]) E. Gott, Electronics **29** (1956) 174.

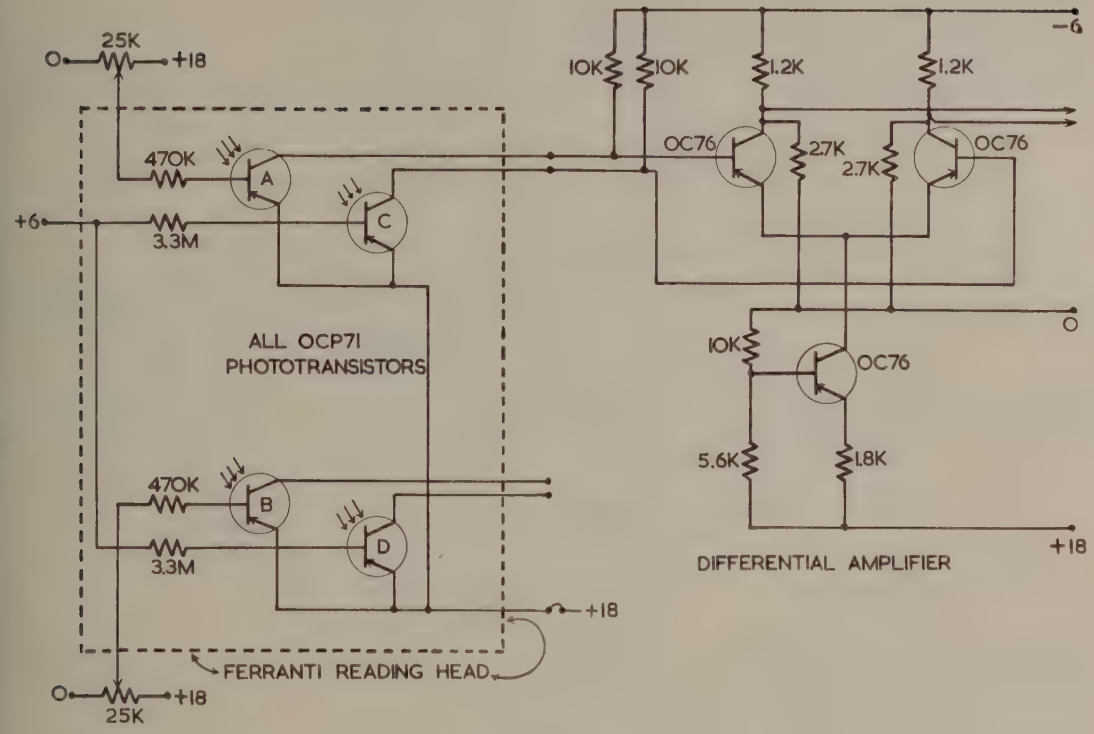

25K
0
+18
470K
A
C
+6
3.3M
ALL OCP71
PHOTOTRANSISTORS
470K
B
D
3.3M
+18
FERRANTI READING HEAD
0
+18
25K
10K
10K
1.2K
1.2K
-6
OC76
2.7K
2.7K
OC76
0
10K
OC76
5.6K
1.8K
+18
DIFFERENTIAL AMPLIFIER

Fig. 4

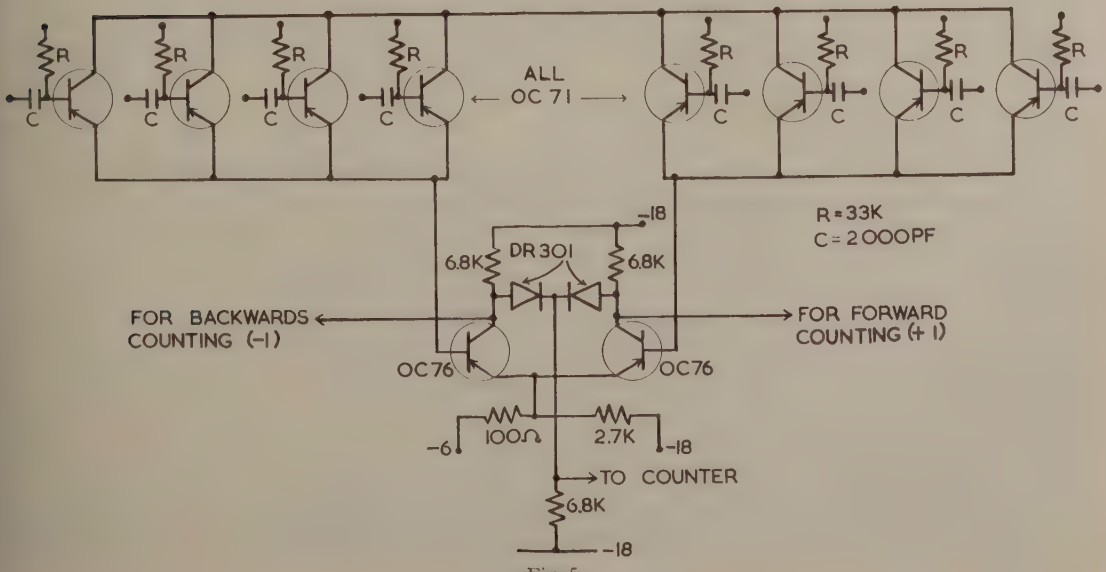

R
C
ALL
OC 71
-18
R = 33K
C = 2000PF
6.8K
DR 301
6.8K
FOR BACKWARDS
COUNTING (-1)
FOR FORWARD
COUNTING (+ 1)
OC76
OC76
-6
100Ω
2.7K
-18
TO COUNTER
6.8K
-18

Fig. 5

1 position. This current will flip the magnetization of the ferrite so that after the copy pulse the instantaneous positions of the binaries are copied down into the ferrites.

output pulse and punch a hole on the tape, and will at the same time be reset. The different capacitors ($C_0, C_1 \ldots C_7$) are charged in series by a ring counter driven by the read out pulses.

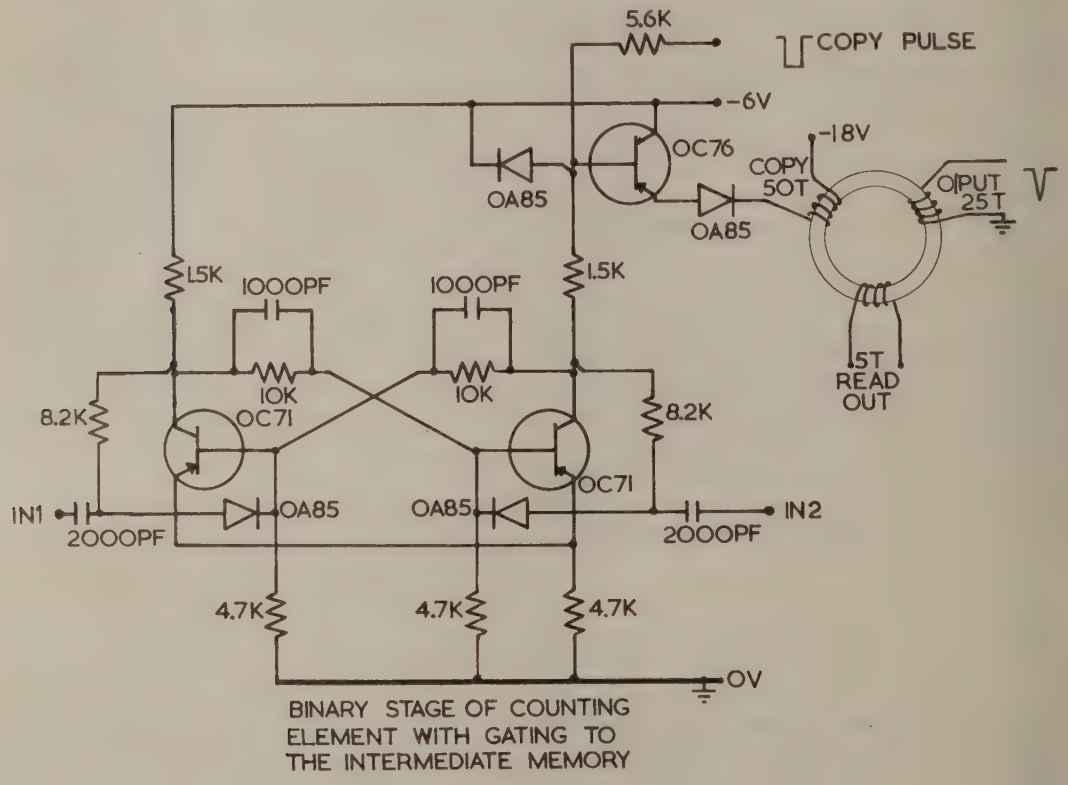


Fig. 6

5.4. READ OUT

The ferrites are read out in 8 groups of 5 beginning with the lower digits of x and finishing with the most significant digits of y. The output windings of a row of 8 cores are connected in series to a flip-flop which actuates one of the 5 coils of the tape punch via a power stage. One 5-core column of the matrix is read out at a time to the tape punch by sending a current pulse obtained by discharging a capacitor through the series connected read out windings of these 5 cores (fig. 7). A core which has changed its magnetization at the copy process and thus stores a 1 digit will produce an

For "on the fly" measurements, a synchronizing unit must be provided to prevent printing out at the time of a count. Appropriate timing circuits control the program for reading out information from the ferrite memory, 5 cores at a time at a rate set by the speed of the tape punch, 25 lines/sec. This gives approximately $\frac{1}{5}$ sec for printing of a set of coordinates.

5.5. COPY PULSE AMPLIFIER (fig. 8)

This amplifier drives the amplifier-gates of the counter chain with a 12V pulse of 20 μs duration.

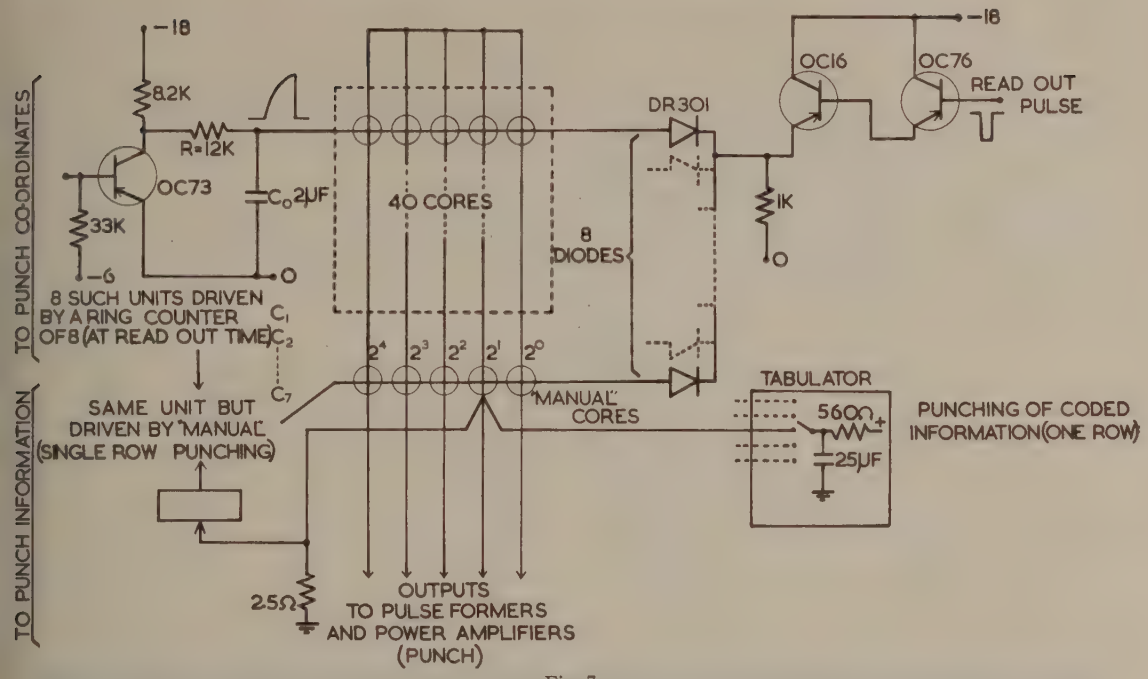

TO PUNCH COORDINATES
TO PUNCH INFORMATION
-18
8.2K
R=12K
OC73
33K
-6
0
C_0 2μF
40 CORES
8 DIODES
DR301
OC16
OC76
READ OUT PULSE
1K
8 SUCH UNITS DRIVEN BY A RING COUNTER OF 8 (AT READ OUT TIME)
C_1
C_2
C_7
2^4 2^3 2^2 2^1 2^0
"MANUAL" CORES
SAME UNIT BUT DRIVEN BY "MANUAL" (SINGLE ROW PUNCHING)
TABULATOR
560Ω
25μF
PUNCHING OF CODED INFORMATION (ONE ROW)
25Ω
OUTPUTS TO PULSE FORMERS AND POWER AMPLIFIERS (PUNCH)

Fig. 7

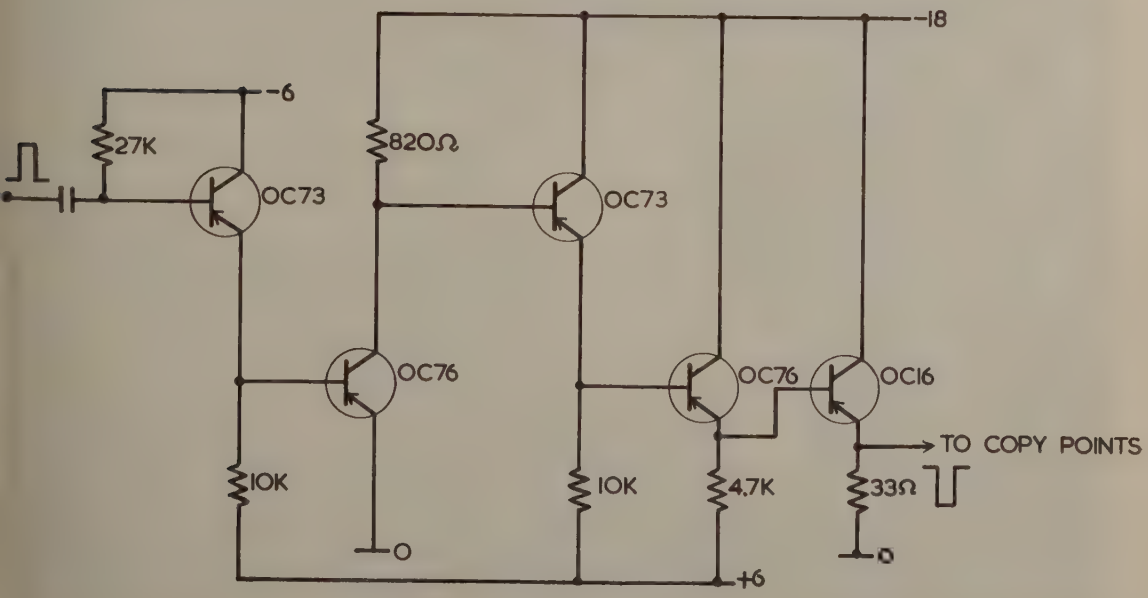

-18
-6
27K
820Ω
OC73
OC73
OC76
OC76
OC16
TO COPY POINTS
10K
10K
4.7K
33Ω
0
+6

Fig. 8

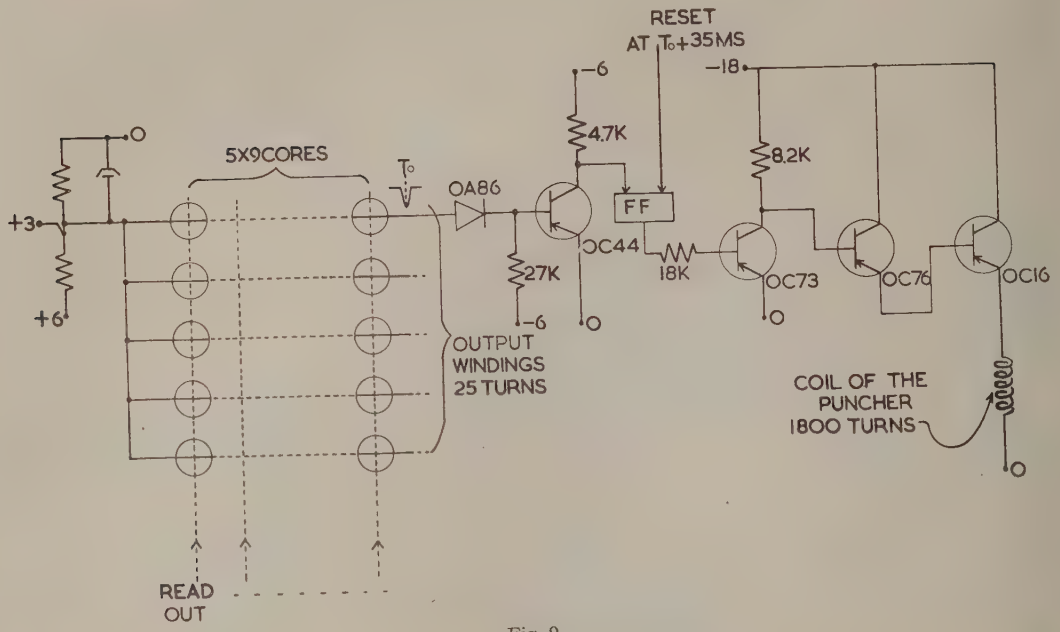

RESET
AT T₀+35MS
-6
-18
4.7K
8.2K
5X9CORES
T₀
OA86
FF
+3
OC44
18K
OC73
OC76
OC16
27K
+6
-6
OUTPUT
WINDINGS
25 TURNS
COIL OF THE
PUNCHER
1800 TURNS
READ
OUT

Fig. 9

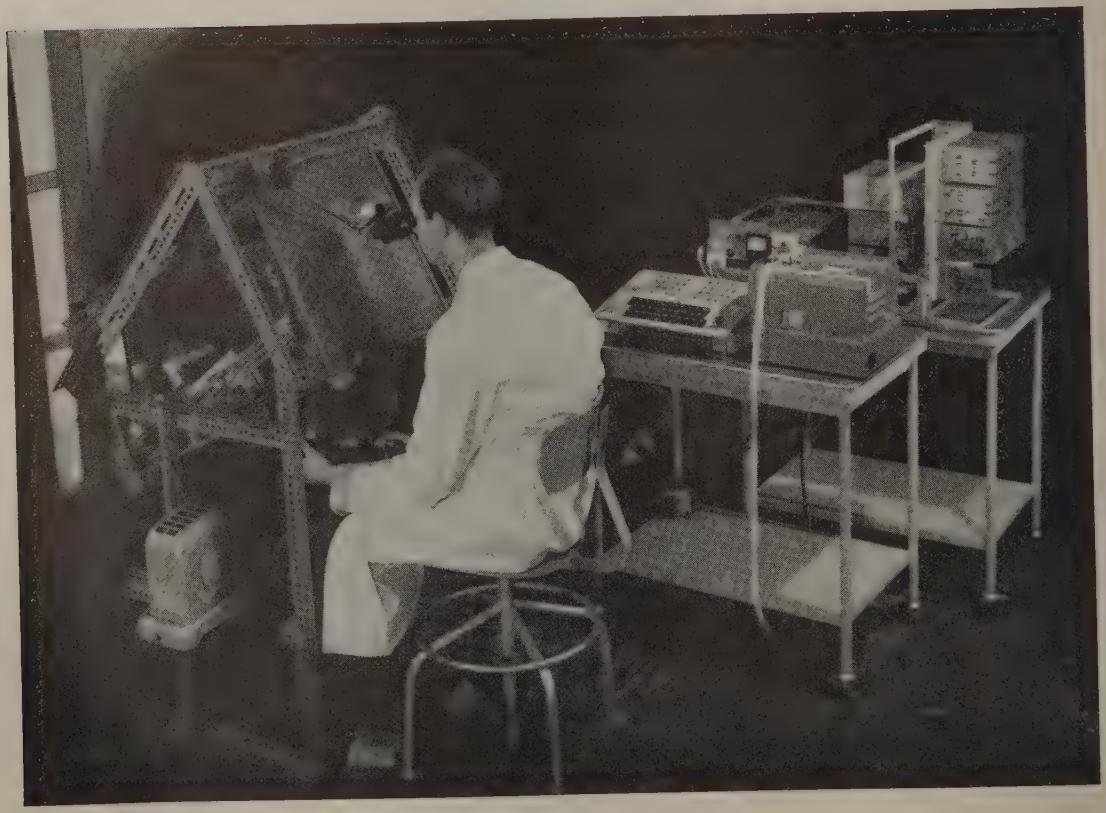


Fig. 10

5.6. OUTPUT SYSTEM (fig. 9)

The pulses from the output winding are negative at read-out time if a hole is to be punched. This signal is shaped, then goes to a flip-flop and amplifier stages to drive the tape puncher.

6. Present State of Development

The electronics is finished and the carriage will shortly be motorised. A manual tabulator has been made for putting in information or test figures (fig. 10). A typewriter will be adapted when it becomes available. Rough tests have been carried out with the closed loop TV system. Linear enlargements of 600 times were obtained with a typical picture. Increased light intensity should bring about a definite improvement in contrast and extend the enlargement factor to 1 000.

DISCUSSION

KANDIAH remarked that it might be simpler and incidentally cheaper to use a flying spot scanner rather than a TV camera to reproduce the area round the crosshair on the projected image.

MEUNIER pointed out that flying spot systems were being used to introduce contrast in pictures lacking in this which might be useful for precision work.

RICHEZ indicated that the life of a flying spot camera would be longer than that of a statical orthicon.

VON DARDEL answering Kandiah and others remarked that a TV camera had been used as it was available and was considered preferable to a simple optical system on the grounds of intensity and ease of scanning. To magnify the crosshair area with say a telescope would impose undue fatigue on the observer and defeat one of the objects of the device.

GOLDSCHMIDT-CLERMONT replying to a question estimated that evaluation of photographs would be about ten times quicker with the IEP than by ordinary techniques.

NUCLEAR INSTRUMENTS 2 (1958) 164; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION III.6

TRANSISTOR CURRENT GENERATOR FOR FAST FERRITE MEMORY MATRIX

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Centro Informazioni Studio Experienze, Milano

A transistorised address current generator has been developed driven by Burroughs MO10R trochotrons. It is used to switch a ferrite core matrix in 1 μs. Fig. 1 shows the circuit which is characterized by high current stability. The resistance stabilized current of 100 mA flowing in T2 and the lower saturated transistor 2N269 is switched into the transformer diode gate load. The second 2N269 is normally cut off and serves to present a high impedance to the gate. As low input currents from the trochotrons are used (6mA) a system of positive feedback has been adopted to aid hard switching.

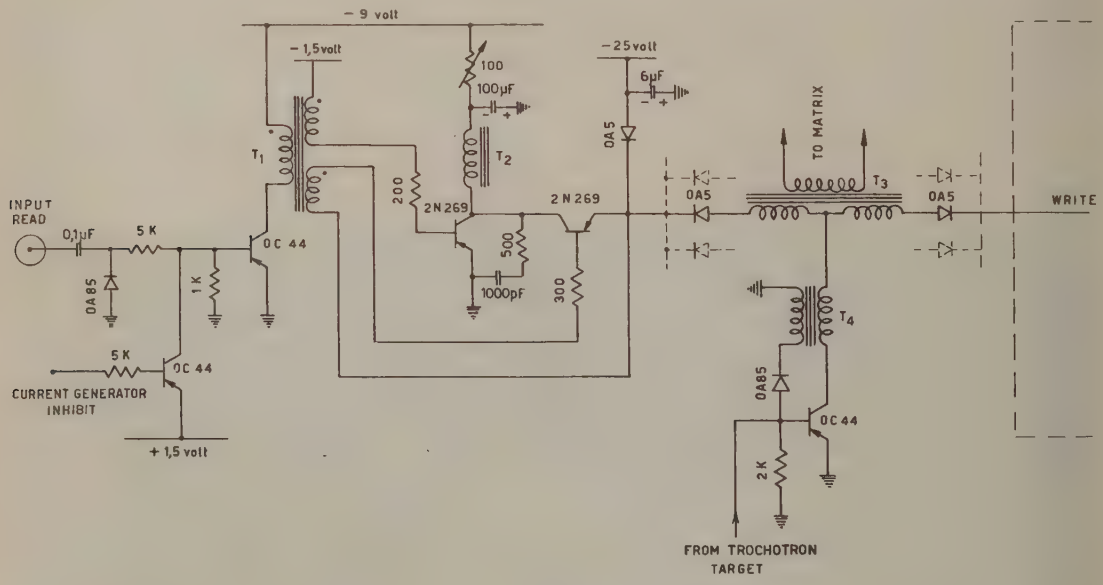


Fig. 1

The circuit has the drawback of a steady current drain.

† C.I.S.E. and Politecnico, Milano.
†† Now at C.N.R.N., Milano.

DISCUSSION

In reply to a question GATTI indicated a step up ratio of 1:4 in the output pulse transformer.

MOODY indicated that similar switching systems had been used at Ottawa in digital computer work and indicated the severe nature of the wattage problems in this new technique of switching core lines with saturated transistors. If there were many cores in series and the transistor is not saturated excessive wattage could build up across it.

NUCLEAR INSTRUMENTS 2 (1958) 165–168; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION III.7

USE OF MAGNETIC TAPES AND CORES FOR NUCLEAR DATA STORAGE AND COMPUTATION

F. H. WELLS

Atomic Energy Research Establishment, Harwell, Didcot, Berks

1. Introduction

The study of the interaction between neutrons and various elements, requires the accurate measurement of neutron energy spectra. The analysis involves sorting neutrons into a large number of energy channels and scaling the number of neutrons arriving in each energy channel during the course of the experiment. Thus a large number of scalers is required for each experiment. As the number of energy channels is increased this collection of scalers becomes uneconomic and is being replaced, in the A.E.R.E. instruments, by a magnetic core storage method supplemented by magnetic tape storage. A typical spectrum showing the need for accurate energy analysis is shown in fig. 1.

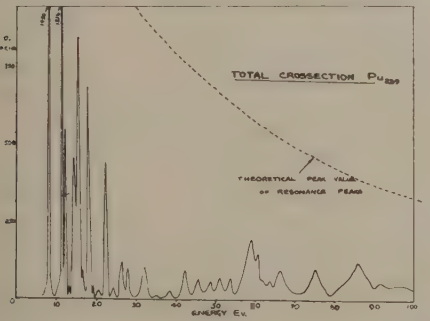


Fig. 1

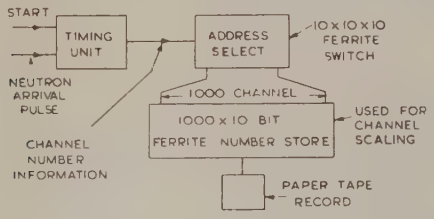
2. 1000 channel neutron velocity spectrometer

This instrument[1]) for the measurement of absorption spectra uses an atomic reactor as neutron source. The neutrons from the reactor pass through a "chopper" or high speed mechanical shutter giving a pulse of neutrons of about 0.5 microseconds duration and maximum repetition rate of 1000 cycles per second. This neutron pulse enters the material under test and the neutrons passing through travel down a long flight tube, up to 100 metres long, and are then detected by a neutron scintillation counter. The time of flight of each neutron thus gives the velocity and hence energy of these neutrons. The neutron detection rate is so low that it is unlikely for more than one neutron to be detected for each burst of incident neutrons. Thus the time between the neutron counter pulse and the preceding "Start" pulse, derived from the time of shutter opening, gives the time of flight of the neutron. This time is measured by the standard technique of allowing a scaler to count pulses from a timing oscillator as soon as the start pulse is received and stopping the scaler counting when the neutron counter pulse arrives. The number registered on the scaler then gives the time of flight of this neutron, in terms of the number of cycles of the timing oscillator. In the instrument being described the oscillator may run at 1, 2 or 4 Mc/s giving time measurement to 1, 0.5 or 0.25 microsecond intervals and the scaler may count up to 1000. Thus 1000 timing or energy channels are available and the number registered on this timing scaler is termed the address of the neutron. This procedure has sorted the neutron into one out of a possible 1000 channels and the number of neutrons arriving in each

[1]) F. H. Wells and J. G. Page, A 1000 channel neutron velocity spectrometer using ferrite data storage, Proc. I. E. E. **104** B, Supplement No. 7 (1957) 470.

of these channels during the course of the experiment must then be scaled. This is performed by the use of square loop ferrite cores.

Fig. 2 shows the block schematic diagram and fig. 3 illustrates the ferrite cores in more detail. The timing scaler mentioned previously consists of 3 trochotron stages giving a scale of 1000 and,



1000 CHANNEL NEUTRON TIME OF FLIGHT SPECTROMETER–TIMING AND SCALING

Fig. 2

when this has stopped counting the timing oscillator, each stage may be considered as a single pole ten position switch, giving a particular address. These switches route suitable currents through windings on one thousand 8 mm diameter square loop ferrite cores in a $10 \times 10 \times 10$ arrangement in such a way that only one core out of the 1000 receives current from all 3 switches. The cores have in addition another winding carrying a D.C. bias and the values of current are such that the magnetising force due to two switch current pulses is roughly equal to that of the D.C. If three switch current pulses are received then the core traverses its hysteresis loop giving current from an output winding. By this triple coincidence current

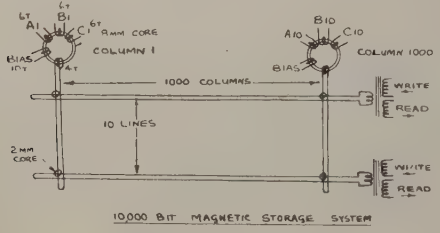


Fig. 3

arrangement only one of the 1000 cores gives an output current which then forms the column current of the number storage matrix, fig. 3.

The number storage in association with a common binary scaler performs the function of 1000 scalers. This storage consists of 10 000 2 mm diameter ferrite cores arranged as 1000 columns each of 10 cores. Each core may be positively or negatively magnetised giving "0" or "1" storage so that each column of 10 cores may store a 10 binary digit number or 1024 neutrons. When a column current pulse is received from the address matrix the number stored in the column is transferred to a transistor binary scaler, one pulse is then added to this scaler and the new number in binary form is rewritten into the column of cores.

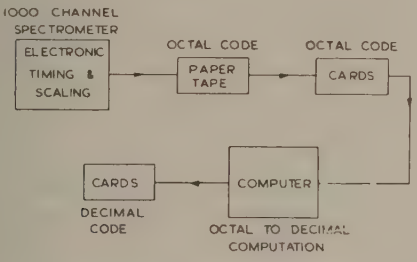
At the end of the counting run, the binary numbers stored in the 1000 columns are read out in sequence and punched in groups of 3 digits at a time on a paper tape.

Each experiment will consist of a counting run with the sample of test material in place and another count without sample. The ratio of these two results for each timing channel will then give the absorption spectrum, after sundry corrections for background counts and total counting periods. Thus simple computations must be performed between two (or sometimes more) sets of 1000 numbers. For this purpose the following system is being installed.

The paper tape record is transferred to Hollerith punched cards the 10 binary code numbers in each channel being punched in a 4 digit octal code form, (group of 3 binary digits being punched as a number from 0 to 7). Five channels are punched on each card. The octal coded numbers are then translated to decimal codes in a computer and a new set of cards punched in decimal code.

The cards with the two sets of experimental results are then sorted and mixed with 1000 blank summary cards and fed into a computer for the main calculation of absorption spectra data. The final output appears on the 1000 summary cards, one card for each timing or energy channel. The results on these cards will then be printed in tabular form or recorded on a graph by an automatic graphical plotter. Figs. 4 and 5 show this sequence of events. The

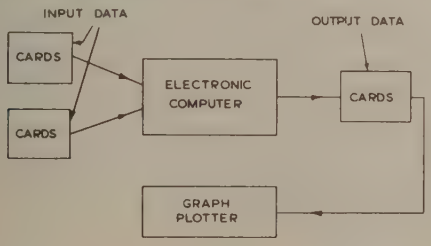
stage of octal to decimal computation shown may be omitted in the future, this calculation being performed during the main computation between sets of cards.



1000 CHANNEL NEUTRON VELOCITY SPECTROMETER OUTPUT DATA RECORDING

Fig. 4

It is desirable to see the results from each counting run before these figures are computed so that incorrect results may be eliminated and to give some guidance as to how much data computation is needed. This requirement will be met by arranging a pen recorder to draw a graph of the results in analogue form during the paper tape punching process.



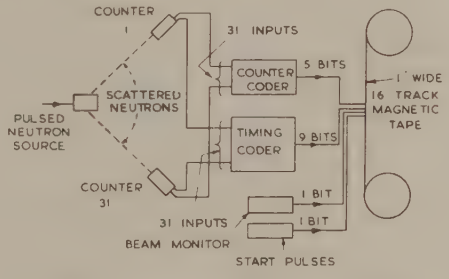
ARRANGEMENT OF COMPUTATION AND GRAPH PLOTTING FROM HOLLERITH PUNCHED CARDS

Fig. 5

3. Neutron scattering spectrometer

In this instrument the neutron source is a "chopper" associated with an atomic reactor as in the previous description but the "chopper" is also designed to select a narrow band of neutron energies. These neutrons enter the material under test and are scattered in various directions. It is required to measure neutron energy spectra at each of a number of scattering angles. The neutron energy is measured by time of flight methods. Since it is necessary to collect as much information as possible from each experiment the instrument will have a capacity of 30 angular positions with about 500 timing channels per angular position. This gives a maximum requirement for 15 000 timing channels into which neutrons may fall and require to be scaled.

This large number of channels cannot be met by conventional ferrite magnetic stores so that magnetic tape recording in conjunction with a small ferrite magnetic store will be used. The principles of this scheme have been worked out by Dr. Egelstaff following the successful use of magnetic tapes for pulse amplitude analysis by Dr. Cavanagh[2]).



NEUTRON TIME OF FLIGHT RECORDING ON MAGNETIC TAPE

Fig. 6

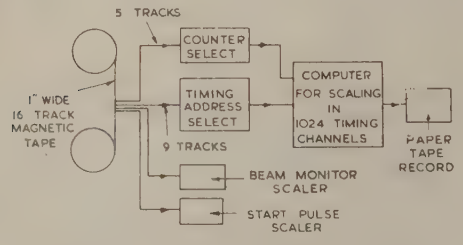
Referring to fig. 6, the time of flight of each neutron will be measured on a 9 stage binary scaler and recorded as a 9 binary digit number on 9 tracks of a 16 track magnetic tape. The path travelled by this neutron will be shown by the particular counter giving the pulse and may be coded as a 5 binary digit number on 5 of the magnetic tape tracks. The remaining two tracks are used for neutron monitoring and recording the number of rotations of the "chopper". Thus all the relevant experimental information is recorded on each detected neutron in the form of a 16 bit code. The neutron counting rates are so low that the tape can be arranged to run slowly at 1 or 2 inches per second. Reels of

[2]) P. E. Cavanagh and D. A. Boyce, Magnetic Recorder for Nuclear Pulses Rev. Sci. Inst. **27** (1956) 1028.

magnetic tape having 7200 feet capacity will be used giving about 20 hours running time per reel.

It should be particularly noted that the electronic complexity involved in this recording process is relatively small so that the reliability of this equipment should be high. This is important since any failure of this recording equipment may lead to waste of Reactor running time. Failures in the reading process described below are not so important as there is no waste of Reactor time.

After this recording process the tape will be read back at a fast rate (probably 50 inches/second) into a 1024 channel computer using a ferrite magnetic store, fig 7. This process will



READER AND COMPUTER FOR NEUTRON TIME OF FLIGHT RECORD

Fig. 7

select neutron information from the tape lying within a preselected batch of 1024 timing channels and scale the number of neutrons per channel in a similar manner to the 1000 channel instrument previously described. After this operation the results of these 1024 channels are recorded on paper tape and then the magnetic tape run through again selecting another batch of 1024 channels. This process is repeated until all possible channels have been read off the magnetic tape and the number of neutrons per channel recorded on paper tape. If the equipment is operating at full capacity then the magnetic tape will need to be scanned 15 times to obtain all the information.

The paper tape record will then be handled by the same system as installed for the previous 1000 channel instrument. This large mass of information will take a considerable time to read from the magnetic tape and compute so that each experimental result may be one or two days running through this data handling process.

4. Conclusion

Magnetic tape is only just beginning to be used in Nuclear Physics Experiments but it gives a possible solution to a variety of experiments where large amounts of data must be taken. The example of neutron spectra is one such experiment where full advantage can be taken of its usefulness. It is very important to keep the electronics in use during the experimental time as simple as possible and this is achieved by the magnetic tape method. One limitation of magnetic tape is its comparatively slow recording speed so that its use may be limited to experiments where the rate of data collection is low.

DISCUSSION

Von Dardel asked if the playback speed was limited by the analysing time of the coding device or was it inherent in the magnetic tape.

Wells replied that the resolution time on the 1000 channel strobing device was much the same as the limitation on the tape speed. The speed of the tape through the reader was limited to 50 in./sec for mechanical reasons. Also a similar limitation existed with the transistor driving circuits on the pick off computer. Pulses closer than 40 μs could not be handled by existent transistor scalers which was in line with 200 pulses/in. on the tape running at 50 in./sec.

Kandiah following up various remarks on magnetic tape indicated that to improve the resolution loss on the 3% observed by Cavanagh would require improvement in the quality of transport mechanisms, feeding heads and general handling of the tape. The use of tape to store video information was acceptable in the entertainment industry where loss of a few microseconds on the tape and non-linearities and short life of the writing head ($\frac{1}{2}$–5 hours) were not troublesome. This was not true for accurate recording work.

NUCLEAR INSTRUMENTS 2 (1958) 169–172; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION III.8

PRINTOUT OF SCALER POSITION

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CERN, *Geneva*

The outcome of a nuclear physics experiment in the form of the number of counts recorded on scalers and mechanical registers is usually recorded manually by the experimenter. In complicated experiments time may be gained and human errors avoided by an automatic recording of the data. Sometimes the experiment is so complicated that faults are likely to occur during a run and it is then of value to have a rapid method of recording intermediate results so that in case of a fault the whole run has not to be discarded.

The simplest solution to the recording problem is the printing mechanical register which however is too slow for many purposes.

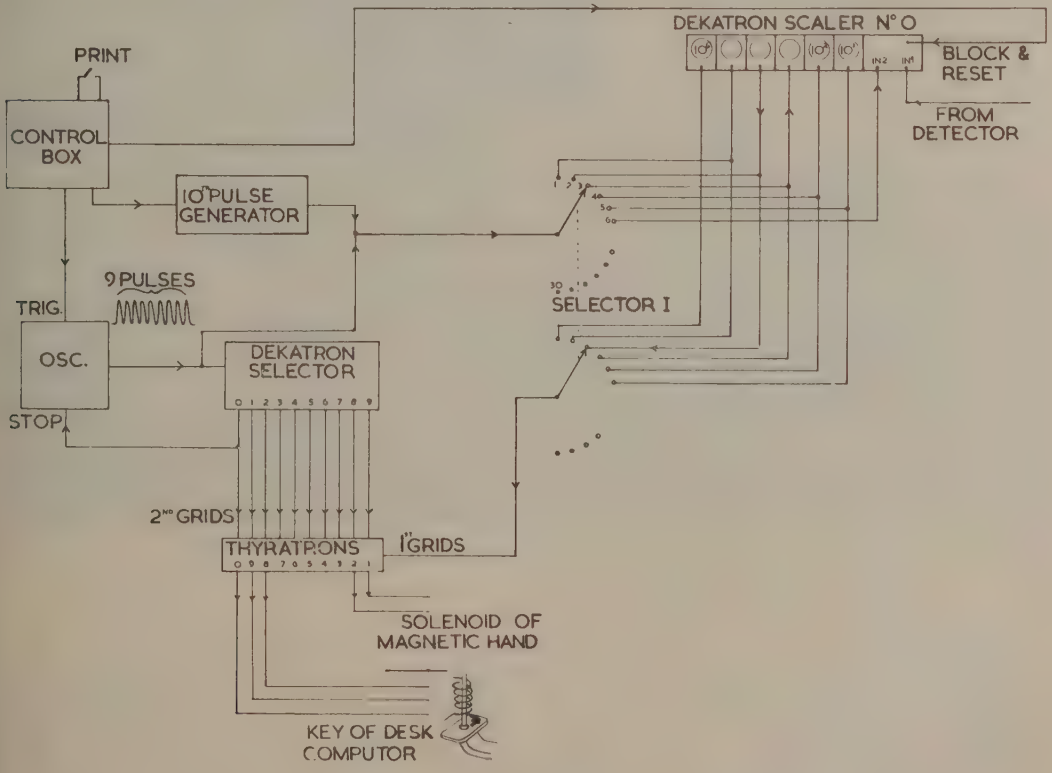

Fig. 1

We may therefore investigate the methods to record scaler positions. In some types of scaler circuits the position of the scaler can be read out electrically by measuring the voltages at suitable points in the circuit. This has been

done for trochotrons in a commercial unit[1]) and for scalers with E1T decade tubes by Gåsström[2]).

For dekatron scalers the position of the scaler is not apparent electrically, except in the position zero. For other positions the position can however be determined by stepping the scaler to the zero position. The necessary number of steps is the complement of the scaler position. A printing unit[3]) for a dekatron scaler based on this principle is being put on the market by a commercial firm. Other types are being developed[4]).

1. Principles of the Instrument

We have developed a print out unit for scalers independently of the previously mentioned instruments. The unit has been designed for operating conventional solenoid operated typewriters or printing desk computers and uses only standard components.

The principle of the apparatus is given in fig. 1 and photographs in figs. 2, 3. An oscillator, gated by a selector dekatron, gives out a train of 9 pulses when triggered by an external pulse or by a manual push button, and then stops. The pulses are fed through one bank of a 30 position telephone selector S_I to the input of one of the scaling stages. The carry pulse from the same scaling stage is fed through another bank on the telephone selector to the first grids of 10 thyratrons. These thyratrons are gated on the second grid by consecutive cathodes of the dekatron selector, and only one of them will fire. The thyratrons operate directly the solenoids of a "magnetic hand" which operates a standard desk pcomuter.

A telephone selector with 30 positions allows five 6-digit scalers to be read and printed in one revolution. If the selector has at least 5 independent banks of contacts and is allowed to make two revolutions per printout the capacity can easily be extended to ten 6-digit scalers by the use of a relay which switches between the two sets of banks for the input and output pulse from the decades.

The printing desk computer used for the printout is of the type having a small 10 button keyboard on which the figures of the number to be printed are first set starting with the first significant non-zero number. Printing and feeding of the paper tape is then effected by pressing the "+", "—" or "≠" key depending on whether the number should be added, substracted or disregarded in the totalling done by the internal memory of the computer. These keys are operated by separate solenoids in the "magnetic hand". Since in general no numerical operations of the simple kinds possible with the desk computer are to be performed on the numbers recorded by the scalers, the printing is usually initiated by the "≠" key. The operation of this key for the required length of time is controlled by the contacts of a second telephone selector II which makes one revolution for every sixth position of selector I. It has been found convenient, particularly when as many as ten scalers are read, to let this selector also program the printing of identification numbers, following immediately after the number recorded by the scaler. This is done by feeding pulses from selector II to the thyratron grids.

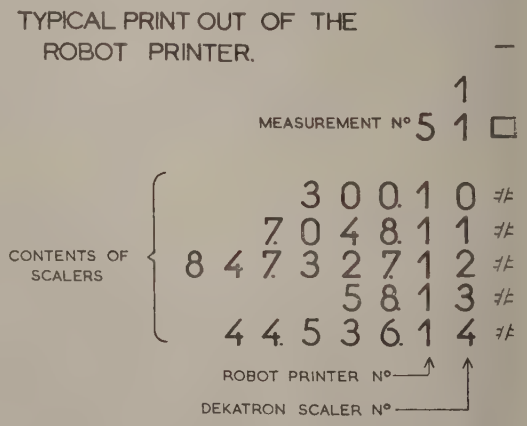
TYPICAL PRINT OUT OF THE
ROBOT PRINTER.
MEASUREMENT N° 5 1 1
CONTENTS OF SCALERS
3 0 0 1 0 ≠
7 0 4 8 1 1 ≠
8 4 7 3 2 7 1 2 ≠
5 8 1 3 ≠
4 4 5 3 6 1 4 ≠
ROBOT PRINTER N°
DEKATRON SCALER N°

Fig. 4

The printout with five 6-digit scalers may thus have the form shown in fig. 4.

[1]) The Robot scaler, made by LKB, Stockholm.
[2]) R. Gåsström, private communication.
[3]) Printing unit produced by Labgear.
[4]) Berkeley-Beckman Instruments, California.

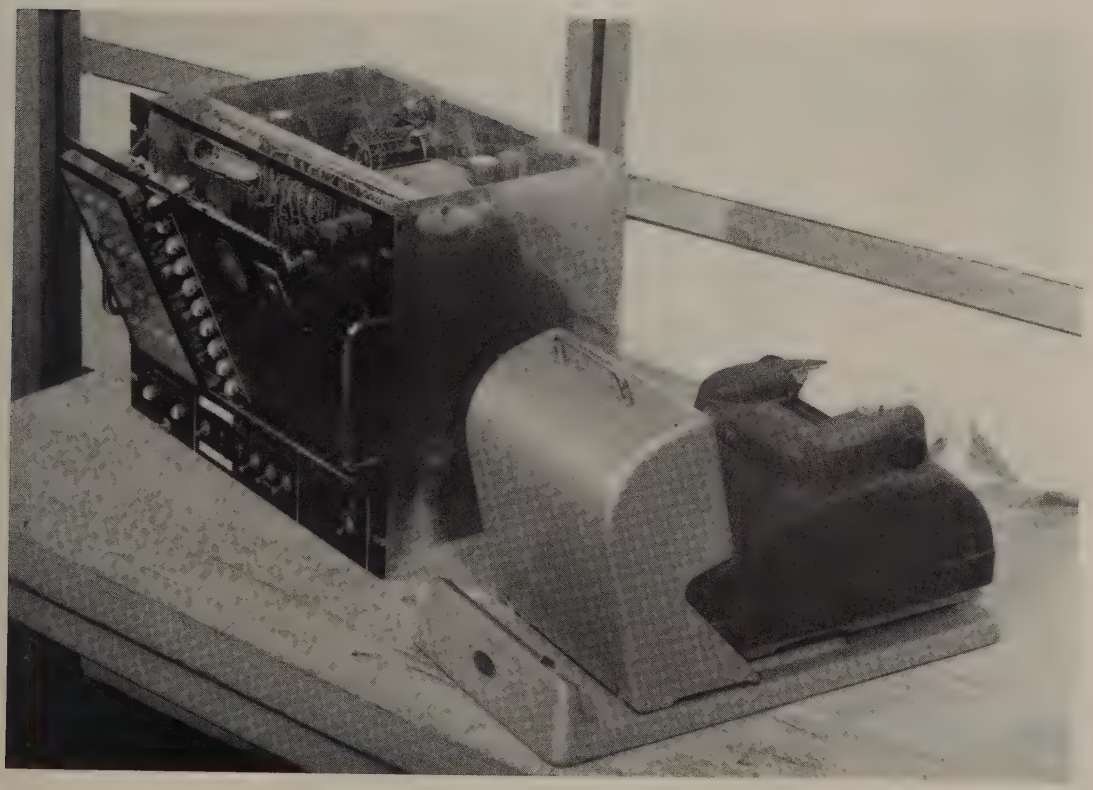

Fig. 2

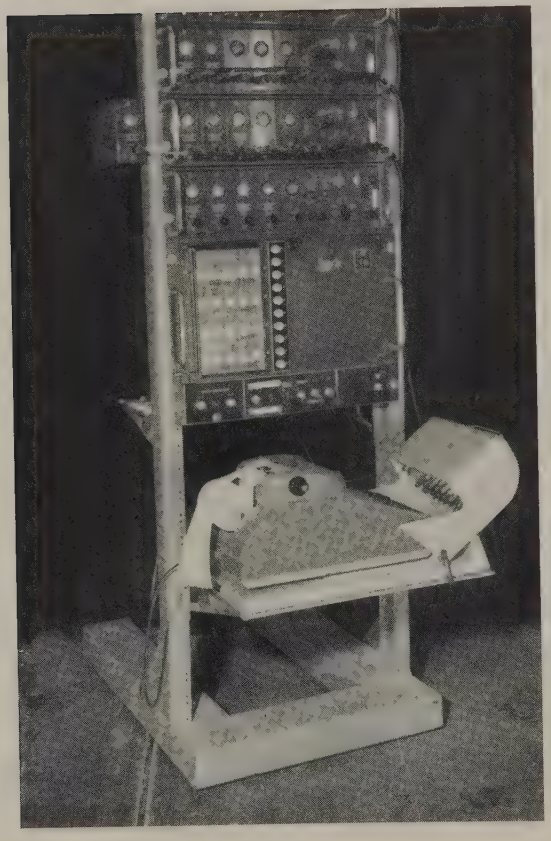

Fig. 3

The first three lines result from an initial program which is intended to separate each printout process from the preceeding and to give it a running indentification number. This identification number is obtained by adding unity each time to the internal memory of the desk computer and printing the accumulated sum. The initial program is obtained in an initial revolution of telephone selector II.

2. Special Features

Since in the process of printout each decade is fed with 9 pulses the number stored in a 6-digit scaler will after printout be increased by 999 999, or more correctly decreased by 1 since there is no decade to record the carry pulse of the first decade. It has been considered desirable that the scaler after printout should be left in its original position. This is important if an intermediate printout is desired during an experimental run after which the scaler continues to accumulate counts without resetting. It is also convenient for testing the proper functioning of the printout unit, by repeated printout processes which are then expected to give identical result. To achieve this desired feature a special relay adds a tenth pulse to the pulse train fed into the least significant decade of each scaler, as shown in fig. 1.

It is essential to block the normal input of the scalers during the printout process since incoming pulses would otherwise interfere with the printout. This requirement generally necessitates the modification of the scaler to provide a special input for the reading pulse train which is not affected by the blocking. The most convenient type of scaler for use with the printout unit is the dekatron scaler with cold cathode valve coupling between stages, since this type is commercially available and requires no modification apart from the above mentioned extra input circuit.

A relay operated after the printout of the last scaler resets all or part of the scalers, depending on the requirements of the experiment. This or other relays of the unit may also be used to change the experimental conditions thus controlling automatically a sequence of experimental runs as well as taking down the results of each run. For such applications the printout process may be triggered by the output pulse of one of the scalers after a predetermined number of pulses.

3. Reliability and Testing

It is obvious that a very high degree of reliability has to be required of the printout unit since otherwise the whole outcome of an experiment will be falsified. The prototype unit so far constructed has been found in about 2000 hours of operation to fulfil this requirement quite well if the instrument is given periodical maintenance particularly as regards some of the relay contacts. It is also important that faulty performance can be quickly detected by a test routine before the instrument is taken into use. One method of testing by repeated printouts of the same scaler position has already been mentioned. In a second test routine the operation of the printout unit is changed in such a way that each decade of the scaler receives a tenth pulse following the normal pulse train of 9 pulses. In this way the number "111110" will be added into the scalers for each printout, and ten repeated printouts should yield the results 000000, 111110, 222220, $\cdots$, 999990. This is a particularly efficient test of the functioning of the desk computer, magnetic hand and associated equipment since all numbers from zero to nine are tried.

4. Speed of Printing

The speed of printing is limited essentially by the speed of the desk computer, which requires 0.1 s to set a figure and about 1.5 s to print a line of figures. These times are longer than the 40 ms length of the pulse train for the read out process. The initial program for giving an identification number of the printout requires about 4 s. The printing of the result of 10 scalers will thus require about 27 s. This time may be reduced somewhat by not printing the zeros preceding the first significant figure of a scaler, and a relay has been incorporated to this effect.

5. Printout of Fast Scalers

The dekatron scalers which have been adapted for use with the printout unit have a resolving power of about 300 μs and have in general to be preceded by a faster scaler. In many cases the position of this prescaler need not be taken into account. In special cases, such as experiments with pulsed accelerators having a low duty ratio it will however be desirable to include this prescaler in the printout. Consideration has been given to the problem of reading their position directly with the printout unit, but it has been found that this will require rather large modifications of the scalers which may interfere with their proper functioning. A more satisfactory procedure seems to be to transfer the counts stored in the prescaler to a dekatron scaler of the normal type which is then read and printed in the ordinary way.

6. Applications

The most important application up to now has been in an experiment on parity where repeated measurements are made with reversal of a magnetic field between successive runs. This experiment is run automatically 24 hours per day, controlled entirely by the printout circuit. The printout circuit is triggered at a predetermined time by one of the scalers which is running on 50 cycles. The input to all scalers are blocked, their positions read and printed, the scalers reset and the input opened to the pulses for the next experimental run. In the meantime a relay actuated by the printer has caused the magnetic field to reverse.

Apart from obvious applications such as this experiment, the printer can be used for storing more detailed information than can be analysed in a conventional multichannel kicksorter. Experiments are considered where the simultaneous pulse heights from say 4 different counters should all be analysed in 100 channels and correlated to each other. The digital information in this case corresponds to 100^4 *i.e.* 10^8 channels. A possible solution to this storage problem is to convert the 4 pulse heights to a train of pulses, using the Wilkinson method for amplitude to pulse number conversion, and to feed these pulses to 4 scales of 100, which are then read and printed with the printer.

While the present prototype model of the printer has performed satisfactorily in all respects, various improvements and simplifications are considered for later models. The use of a solenoid operated typewriter, although more expensive than the desk computer, would greatly simplify the subprogram. The printout time would be decreased from about 25 seconds to 7 seconds for printing ten 6-digit scalers. The presentation of the data in the tabular form provided by the typewriter is more convenient than the long paper strip of the desk computer. It is also envisaged to replace the majority of the relays by transistor circuits. The telephone selectors could possible be replaced by a multiway electronic switch, using cold cathode triodes or transistors and diodes.

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SESSION IV.1

SYSTEM RELIABILITY

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The question of the reliability of electronic equipment has attracted more and more attention as the users of to-day's complex systems have started to ask for dependable daily service from their equipment rather than just occasional demonstrations.

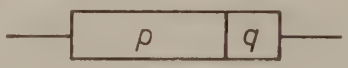
Several big investigations have concerned themselves with the unreliability of military equipment and components, and much has been learned about how and why they fail—information which is now being used by manufacturers to improve their products.

The components which we use in larger and larger numbers will, however, never become perfect, which means we will have to face the fact that our electronics will fail. Fortunately that risk can be minimized, and we will now look at the ways in which this can be done.

Components and circuits fail for two major reasons: they *wear out* or they fail *catastrophically*. Wear-out failures can—at least in principle—be anticipated by measuring some electrical properties, and the component can be replaced before the circuit in which it works becomes inoperative, but there is no way of foreseeing catastrophic failures; one only knows that they will occur, but not when.

We will, in the following, concern ourselves mainly with catastrophic failures and, although accurate information regarding components' failure risks and general behaviour is very difficult to obtain, one can draw some conclusions of practical interest even from rather crude models.

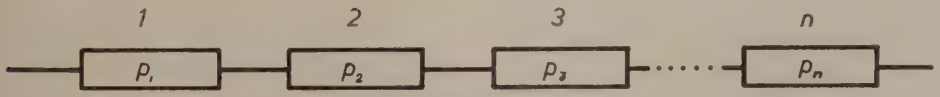
We will mean by the *reliability* (p) of a component its chance to work satisfactorily for a given time. The *unreliability* (q) is then the chance that the component will fail during that same time, and the sum of p and q equals 1, or certainty (fig. 1).



p = reliability
q = unreliability
$p + q = 1$ (100 %)

Fig. 1. An element.

If one connects a number of components, or *elements*, functionally in series to form a *system*,



$P = p_1 \cdot p_2 \cdot p_3 \cdots p_n$

or: $P = p^n$ (*with identical elements*)

Fig. 2. A series system and its reliability.

the reliability of that system becomes the products of the reliabilities of the individual elements or, for identical elements, p^n (fig. 2). This is the unfortunate arrangement which we generally use, and the only way of making it perfectly reliable seems to be to use perfect components—or none at all.

Another way of connecting elements into a system is to use them functionally in parallel; that is to say to let one element do the job alone, but if it fails to make a new element take over, either by being switched in, or by actually

working in parallel all the time (fig. 3). The *un*reliability (Q) of that combination then becomes the product of the *un*reliabilities of each element, or, for equal elements: q^m, where m is the number of parallel elements. The system reliability P now becomes $1-Q$, and here we see that the reliability tends toward 1 for an

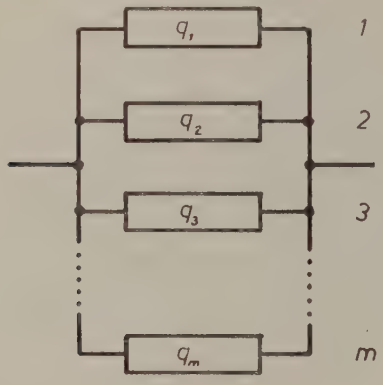


Fig. 3. A parallel system and its reliability.

infinite number of standbys—a useful property of this arrangement.

We can try to use this property to increase the reliability of an existing series system by adding similar systems in parallel, so that, if one system fails, a new one takes over (system standby). The expression of the reliability of such an arrangement becomes

$$P \quad 1 \quad (1 \quad p^n)^m$$

(using identical elements for the sake of simplicity), and fig. 4 shows how the reliability increases with an increasing number of standbys. It is seen, however, that the reliability still tends towards 0 for an infinite number of series elements even if the number of standby systems becomes infinitely large, a fact which indicates that this solution is not ideal.

There remains one possibility and that is to parallel each element before the series connection is made (element standby) (fig. 5). That gives the reliability expression:

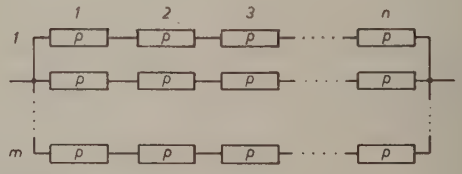
or

$$P \quad (1 \quad q^m)^n$$
$$P = [1 - (1 - p)^m]^n$$

and here we see that the reliability tends towards perfection if we have an infinite number of standbys, even if the chain becomes infinitely long. To use an infinite number of standbys is, of course, not practical, but a similar effect, is in fact, very nearly obtained by using only one spare, if no faulty unit is allowed to remain unrepaired in the system.

Up to now it has been assumed that a new element or system can be switched in with perfect reliability. This is obviously not true, but it can be shown that if the switching device has a reliability which is at least equal to the unit it is there to supervise, element standby is still better than system standby; this in spite of the fact that in the latter case only one switch is required.

SYSTEM STANDBY



$$P = 1 - (1 - p^n)^m$$

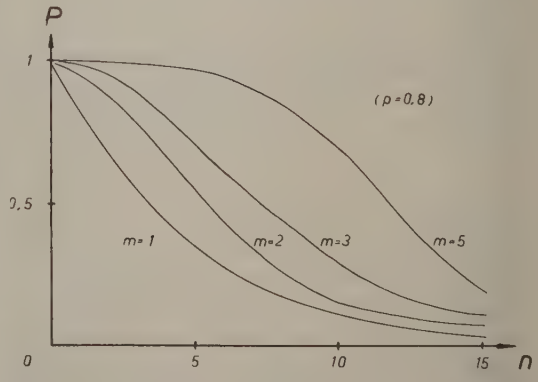


Fig. 4. The moderate increase of reliability when using system standby.

From these simple models one can draw some general conclusions. For the highest reliability, the basic system should be simple, and use good components. If that is not good enough, one

can apply the spare wheel principle and use redundant circuitry, in which case one should try and break up the system into as small units as possible before parallelling. In this way one can obtain any desired degree of reliability with unreliable components. This is, in fact, the secret behind the reliability of such highly complicated systems as telephone networks, even if their redundancy has got other reasons for it than reliability.

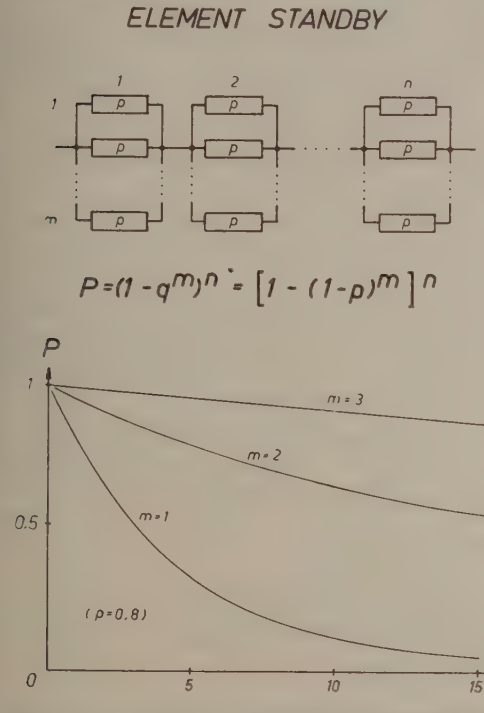


Fig. 5. The high increase of reliability with element standby.

For the designer it would be very desirable to be able to predict the reliability of a piece of equipment before it is actually made. This is very difficult since accurate reliability figures can only be obtained from life tests on a large number of the apparatus itself. Some rough estimates can, however, be obtained by using existing life test data on some of its more important components.

Having a large number of identical components on life test one can note how many of the components fail for each hour. The number of components failing per hour divided by the number of survivors is a measure of the risk each component runs of failing catastrophically during that hour, and is called the *hazard* (z) (fig. 6). The curve obtained, which is very

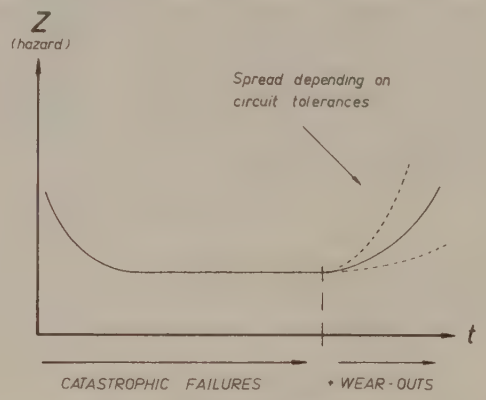


Fig. 6. A fundamental hazard curve.

general in nature and applies equally well to engineers as to valves, shows a high risk for early failures, a constant hazard over a considerable time and increasing risk as the component wears out. (From various investigations one will, for example, get hazard figures for minature valves ranging from 1/4000 to 1/500 000 and for resistors and capacitors from 1/50 000 to $1/10^7$.) The region of constant hazard indicates that a population of components decays exponentially during that time. The percentage of survivors at any given time is also the chance any one component stands of surviving that period of service, or its reliability. The expression for component reliability can then be written

$$p = e^{-zt} .$$

For a system using different components, the over-all hazard becomes the sum of the individual hazards, and the system reliability for a certain period of operation can then easily be computed.

Another useful measure of the reliability is the time (T) at which P is down to 1/e or 37 %, which is called the *mean time to failure* and is $1/z$.

As an example we can calculate the reliability of a piece of equipment containing 250 valves

and 2500 other component parts, using the conservative hazard figures of 10^{-5} for valves and 10^{-6} for the other components. This means we are using first-class long-life valves in sockets that really fit, with well regulated heaters and at low electrode voltages; that our resistors and capacitors are running at optimum ratings, neither too high nor too low, and that the temperature is comfortably low. The hazard becomes:

$$250 \times 10^{-5} + 2500 \times 10^{-6} = 5 \times 10^{-3}$$

$$T = \frac{1}{z} = \frac{1000}{5} = 200 \text{ hours} .$$

That is to say, we have a 37% chance that the apparatus will work for 200 hours. If we only want to use it for 10 hours we get

$$P = e^{-\frac{10}{200}} = 0.95$$

or 95%, which is excellent, but we will have less than 1% chance to operate it for as much as 1000 hours without a failure.

If this is not good enough for us, we will have to use standby equipment, but the whole system should then preferably be engineered for this technique right from the beginning, since demands on components, lay-outs and testing facilities may be quite different.

But before resorting to redundancy it may well pay to try and really find out what the equipment is there to do, and perhaps find a simpler and thus more reliable way of doing it.

This leads to some points which should be considered when designing any form of complicated system where reliability is important and down-time expensive.

(1) Realistic specifications.

This will allow the simplest solution to be chosen that satisfies the system, and keeps unnecessary refinements to a minimum.

(2) Mature, conservative circuit design.

(3) Good components, or as good as the budget will allow.

(4) Redundancy where necessary.

(5) Maintenance and service eased by simple built-in testing facilities, unitized construction and use of standard parts.

(6) Simple and logical arrangement of controls and displays.

This last point is important, since, from a reliability point of view, it is immaterial if the apparatus stops because of á component failing, or because of the operator accidentally pressing the wrong button, which was improperly labeled, placed in the wrong position or should not have been there at all.

So when designing complicated equipment for reliability one must not forget one very unreliable element—the operator.

References

IRE Proc. on Reliability and Quality Control. Sept. 1956.

K. Henney, Reliability Factors for Ground Electronic Equipment.

DISCUSSION

KANDIAH opened the discussion with a few general comments on reliability. Quantitative studies on this question in six or seven years had accumulated a lot of information and in particular the nuisance values of many types of component were known. The important step in designing new equipment was that the designer avoid most heartily those components with proven bad properties i.e. the ones known to be subject to catastrophic failure and choose components that wear out only very slowly. Having done this he could then build in marginal checking facilities to look after the unreliability. Some of the arguments expressed by Sagnell to use series and parallel systems were difficult to apply to equipment intended for physics use e.g. time of flight spectrometers and pulse height analyzers were instruments difficult of subdivision and the standby technique was considered more applicable to computer devices which used generally a large number of similar elements.

SAGNELL opined that the stacked discriminator was the most reliable type of pulse height analyzer possible. 50 channels stacked in voltage and feeding mechanical registers had the advantage that a failure in any one channel left the remaining 49 in order and the count already in the failed channel was still available. Scalers could be inserted in each channel to improve the resolution.

KANDIAH retorted that such a device is inherently slow and suffers from the serious drawback that there is no

check on mutual drifts between channel boundaries during an actual experiment. The more complicated analyzers had been designed to eliminate this disadvantage which was very important to experimenters looking for the significance of a very minor jump on a spectrum.

Scarrott pointed out the difficulty of eliminating the human element in complex systems. Faults were made during mechanical assembly that often were difficult to trace e.g. dry soldered joints might not be discovered until the equipment had seen years of use. A wider use of automatic assembly methods was urged.

Kandiah concurring said that excellent design work could be made entirely unreliable by bad assembly.

Hutchinson pointing out that the reliability of many things increased if they were kept cool, strongly urged designers to avoid having horizontal metal surfaces in electronic equipment. The best way to ensure adequate ventilation was to mount chassis vertically and provide air flow vents.

Gatti brought notice to a failure of wire wound resistors in a particular equipment. These were conservatively rated but neighbouring components had heated up the ceramic body which on expansion had broken the fine wire.

Kandiah pointed out that it was common fallacy to believe that reliability increased if a component was derated. This was not always the case and for many components such as electrolytic capacitors and certain kinds of resistor there was an *optimum rating*. Experience was showing that electronic equipment was more reliable if aged or run in. Vintage stuff (but not very old) was better.

NUCLEAR INSTRUMENTS 2 (1958) 178–181; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION IV.2

APPLICATIONS OF ISOLATED POWER SUPPLIES TO INSTRUMENTATION PROBLEMS

JOHN H. REAVES and JOHN F. WALTON

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There are several important problems common to many different instrumentation applications that have a possible solution by techniques employing isolated power supplies. A brief description of this type of power supply will be given in this paper, together with the reasons for its use, and two examples of typical applications will be described. More complete information on power supplies of this type and their applications may be obtained from references [1,2]).

Conventionally, D.C. power supplies are employed with one of the two output terminals grounded or by-passed with a capacity to ground. In contrast, an isolated power supply is defined here as one that is designed to be employed with neither output terminal grounded or by-passed to ground. Such a supply is constructed by using a special power transformer having the secondary winding separated from the core and primary by an air gap, which gives a low value of shunt capacitance to ground. The rectifiers and filter circuit elements are also mounted in such a manner as to contribute only a small amount of capacitance to ground. The total shunt capacitance can thus be kept quite small, in practice about 20 mmF. With such a low value, the power supply can be employed in the unique manner in which it is intended, without appreciable deterioration of the rise and fall times in high-speed circuits.

The basic circuit that illustrates the use of an isolated power supply is shown in fig. 1. The power supply is labeled by the trade-name ISOPLY, which is an abbreviated form for isolated power supply. Although a transistor has been chosen for this illustration, a vacuum tube may of course be used in the same way. As shown, the power supply is connected between the collector and the grounded load resistor R_L, whereas in conventional practice, the power supply is connected between the load resistor and ground.

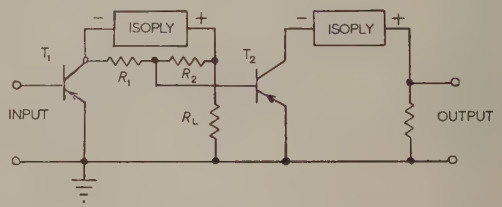


Fig. 1

The circuit of fig. 1 has several important advantages over the conventional circuit. The first is that the proper D.C. level for direct coupling to the base of the following transistor is obtained without any alteration of the signal amplitude. The average or D.C. level of the output signal with respect to ground can be adjusted to be zero, positive, or negative by proper choice of R_1 and R_2. For any of these cases the output signal amplitude remains the same and is equal to the full signal voltage obtained across R_L. In the interesting case when the second transistor is of the opposite type from the first, that is, an NPN instead of the PNP type, the resistors R_1 and R_2 may be omitted and the base of second transistor

[1]) J. H. Reaves, Electronics (McGraw-Hill Publishing Co., New York, 1954), Vol. 27, No. 8, pp. 172 and 173.

[2]) Bulletin No. 1–1256 published by Elcor, Incorporated, McLean, Va.

[5]) Electronic Design (Hayden Publishing Co., Inc., 1957) p. 34.

directly connected to the ungrounded end of the load resistor.

A second advantage of the circuit of fig. 1 is that its stability is much better than that of the conventional circuit. Fig. 2 shows why this is true. The common emitter characteristics of a transistor resemble the anode characteristics of a pentode, in that the current is primarily dependent on the input signal and relatively independent of the collector (or anode) voltage. This causes the output signal amplitude v_0, obtained as shown by drawing the load line, to

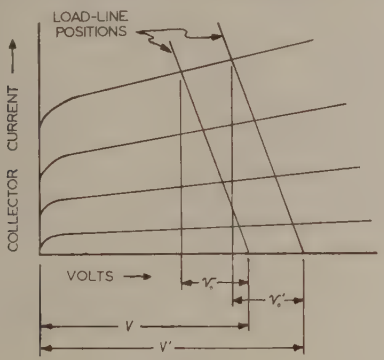


Fig. 2

be independent of the supply voltage, at least to the extent that the characteristic lines are horizontal. The output voltage depends solely on the current, which in turn essentially depends only on the input base current.

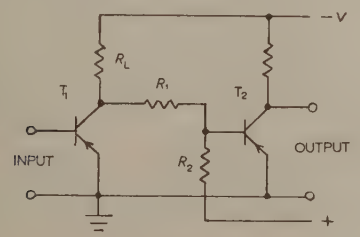


Fig. 3

In contrast, the conventional circuit, which is shown in fig. 3, has an output voltage that depends not only on the current through the transistor, but also on the supply voltage. Assuming that the collector current is constant with respect to collector voltage, the voltage drop across the load resistor will be constant, and therefore any change in supply voltage will appear at the collector of the first transistor as a change of the same value. For example, a two-volt change in supply voltage will result in a two-volt change in the potential between the collector of the first transistor and ground. If the resistors R_1 and R_2 in the attenuator are equal, the effective voltage change transferred to the base of following transistor (in the Thévenin equivalent circuit) will be one volt. This is a far greater change than would be produced in the case with the circuit of fig. 1.

In the case when a triode is used instead of a pentode or transistor in the circuit of fig. 1, the output voltage changes slightly with supply voltage, but not nearly so much as in the case of the conventional circuit. The reason for this is that with the conventional circuit a given percentage change of supply voltage will result in a greater percentage change of the voltage coupled to the following stage. As an example, consider a supply voltage of 150 volts with a variation of 1% or 1½ volts. Since the load resistor is small, the average value of the voltage at the plate of the triode will be in the neighborhood of 140 volts and have a fluctuation of about 1.4 volts. The signal voltage, which is about 10 volts is direct-coupled to the following stage through the voltage divider (see fig. 2). The divider is assumed to have an attenuation of two, and both the signal and the undesired fluctuations are attenuated by this factor. This voltage divider lowers the average potential from approximately 140 volts positive at the anode of the triode to something of the order of five volts negative at the grid of the following stage. The fluctuations at the input of this stage due to supply voltage changes will be about 0.7 volt, which now amounts to 14%. With the ISOPLY circuit, however, the output fluctuations will be roughly equal (on a percentage basis) to the power supply fluctuations, or about 1%.

As already explained, even less fluctuation is obtained using a transistor in the circuit of fig. 1 because of the relative independence of the collecter current with changes of collector supply

voltage. Thus, the comparison of the conventional circuit and the ISOPLY circuit with respect to stability is even more favorable to the ISOPLY circuit than in the example just discussed, which was based on a triode vacuum tube.

Using the conventional voltage divider method of direct coupling, a difference amplifier circuit must be employed if the stability inherent in itself more than compensates for the slight additional cost of the power supplies.

A wide-band direct-coupled amplifier that employs two isolated power supplies in a feedback circuit is shown in fig. 4. With the component values shown the frequency response extends up to about five megacycles, and the maximum amplitude is approximately twenty volts. The use of this circuit as a pulse amplifier

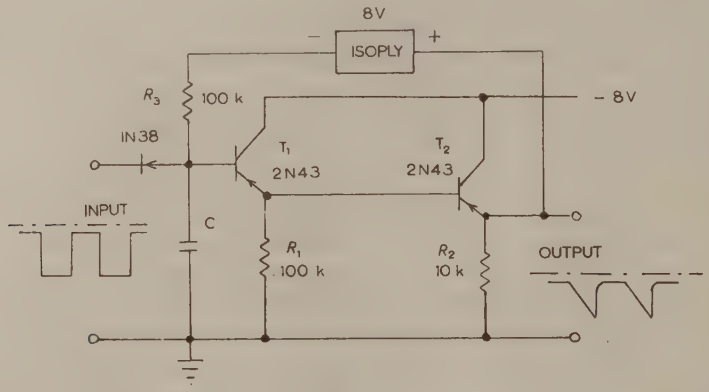


Fig. 4

circuits using isolated power supplies is to be approached. The added complexity and additional power requirements of the difference amplifier circuit are factors that must be considered in any comparison of relative costs.

An objection to the circuit of fig. 1 that is frequently raised is that the cost is greater than for conventional circuitry because many separate power supplies are required. However, if one calculates the total cost of the power supplies required with conventional circuits, and divides this by the number of separate stages that operate from these common power supplies, the cost is found to be surprizingly great. With high-speed circuits, where the current per stage is necessarily high, this cost will be of the order of $ 35.00 per stage. To accomplish the same result with circuits employing isolated power supplies, an expenditure of twice this amount for the power supplies may be required. However, the economy afforded by the extreme simplicity of the basic circuitry for nuclear instrumentation is being considered because of its good linearity and non-overloading characteristics.

Another example of the use of an isolated power supply is shown in fig. 5. This bootstrap

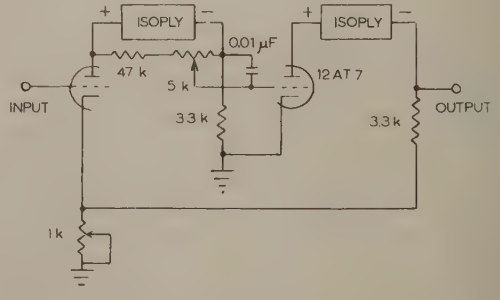


Fig. 5

sweep circuit employs two emitter followers in cascade, and because of the near-unity gain of the emitter follower circuit, the output ramp voltage has good linearity. The circuit also has the advantage that the rate of rise of the ramp

voltage depends on only one voltage, a resistance, and a capacitor, all of which can be held quite constant when high precision is required.

Many other uses of isolated power supplies are being investigated, and the circuits that have been given serve mainly as examples of how such power supplies can be used. The use of ISOPLYS in unitized or modular equipment appears particularly advantageous because of the two additional advantages of convenience and non-interaction afforded by the use of individual power supplies in modular design.

DISCUSSION

After Reaves' paper, MAEDER and MOODY opened the discussion by pointing out the alternative methods of compensating for the shift in DC voltage from input to output in a stage of a DC-amplifier.

MOODY pointed out that if the amplifiers are fed from a positive and negative supply and the coupling between the stages is done by a potentiometer in such a way that the successive grids (or bases) are at ground potential, then the dependence on power supply voltage will balance out if the current in the active devices is proportional to the power supply voltage, a condition which is easy to arrange. This technique has been used in computer work, but has the disadvantage, as

MAEDER pointed out, that the gain per stage is decreased by the attenuation of the coupling potentiometer. This loss in gain may however be avoided by the use of Zener diodes as a constant voltage coupling element between stages. The disadvantage of the Zener diode, i.e. its high temperature coefficient, can be minimized by designing the circuit so that the diode works at an optimum voltage, around $6 \div 8$ volt where the temperature coefficient is small or zero.

Another method to get around the problem of the DC shift from input to output is to use alternatively pnp and npn transistors.

MOODY reported that he had actually used this method.

REAVES pointed out that circuits methods where two large voltages are balanced out tend to be unstable. Moody conceded that this is true if one of the voltages is fixed and the other moves but not if both vary by the same relative amount.

WELLS inquired whether a well stabilized power supply is now not such a standard item that one could rely on its stability. Reeves was of the opinion that it may be better to design the circuit in such a way that power supply variations are not serious, rather than to rely on the stability of the power supply.

The actual construction of Reaves' plug-in power supply was then discussed. The transformer is a regular transformer except for an air gap between core and the secondary winding to decrease the capacity. Rectification is done by a bridge circuit using silicon power diode and stabilization by a voltage regulator tube or for lower voltages by a zener diode. The most essential point of the power supply is a metal shield which surrounds the rectifiers and the secondary winding, and is connected to the output terminal of the power supply. Without this shield too much hum is induced and spikes occurring at the mains frequency are obtained from the charging and discharging of the stray capacitance of the secondary winding. This hum has probably discouraged previous attempts to use floating power supplies in high impedance applications.

MAEDER suggested that the advantage of the small isolated power supplies had to be weighed against the lower stability compared to a big common power supply. Reaves considered that by judicious design of the circuit as indicated in his contribution the circuit was not very sensitive to the stability of the power supply. In addition there is no interaction between different parts of the circuit since each part has its own power supply.

MAEDER pointed out that in transistor amplifiers with a common power supply one usually has a feedback for the base bias so that each transistor adjusts itself and that the method isolated power supplies where this feedback is absent might lead to difficulties. He indicated that the isolated power supply might be useful to make constant current source drawing current from a negative point, a problem for which he knew of no other solution. Moody pointed out that this can also be done with pnp transistors.

NUCLEAR INSTRUMENTS 2 (1958) 182–192; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION V.1

THE PRESENT STATE OF THE TRANSISTOR AND ITS ASSOCIATED CIRCUIT ART

(Text of a spoken paper)

N. F. MOODY

Defence Research Telecommunications Establishment, Ottawa

1. Introduction

I hope that some of you who are gathered here to-day are intending to design an electronic instrument. If you are, you will be faced with a major decision. This decision will dictate not only the form of the circuits you will use, but indeed it will control the very operations which these circuits are to perform.

This decision which you must make is the basis of my talk today. It is this: shall I use transistors or vacuum tubes?

Now at the outset I must tell you that the transistor is a nuisance. It is different from the vacuum tube in all important respects. It is true that both devices have one trivial property in common, for both give power amplification. To designers who are familiar with vacuum tubes, this trivial resemblance between the devices has been a trap. Many of their failures are strewn along the path of progress.

The transistor is a nuisance because you must learn an entirely new circuit technique to use it satisfactorily. Even then, you must avoid a tube-by-tube replacement by transistors. You must re-design on a functional basis. If you wish to use transistors to advantage, you must first work with them for at least a year, for it will take this period of time to master the new techniques. Indeed it will take this time for you to master the transistor itself!

Is all this work worth while? The answer is yes!

The popular articles will have told you the eye-catching advantages of transistors. I should like to add two of my own. First of all, correct design will yield a reliability of operation which tubes cannot approach (and incorrect design will cause the most dismal results). Secondly you have a new freedom to make circuit inventions. For while it is true that the transistor cannot at present replace all vacuum tubes, it is equally true that the transistor can do things which are impossible for the vacuum tube.

How often has the tube designer wished for a device which could work with positive charge carriers rather than electrons? The transistor designer has both types of device at his choice.

In an introductory talk of this nature I cannot develop transistor circuit techniques for you. In fact, I shall not have time to give circuit details at all. Perhaps I can discuss practical circuits afterwards with those who have a detailed knowledge of the field.

I shall try, instead, to contrast the properties of the most modern transistors with each other and with those of vacuum tubes. Thus I hope to interest both the advanced worker and the newcomer. In the process, a few of the basic design approaches will develop themselves. I hope to start the tube people thinking about transistors, and at least direct them to the beginning of a path which will lead to safe transistor circuit design.

We cannot accomplish good circuit design until we know something of the properties of the transistor itself. This alone is quite a study, so I have chosen an unorthodox method of approach. My approach is non-rigorous but it does emphasise the important design features. I believe that Prof. Baldinger will supply some of the missing rigour!

The major properties of the transistor are dominated by two factors. One of these is the

concept that transistor action is based upon rectifier type junctions, and so can be properly represented by semiconductor diodes. This concept is so important in pulse circuitry that I must emphasise it. The other is that a transistor has a built-in time delay to all signals which traverse it. We shall describe the transistor in terms of these properties, starting with the rectifying properties of junctions.

My first concept is that transistor junctions are like semiconductor diodes. If a transistor junction is a diode, why not draw it as one? In fact this is what I shall do, for the Canadian transistor symbols are based on the diode. It is proper, and quite possible, to develop many transistor properties from simple diode concepts, and I shall now show you how this is done.

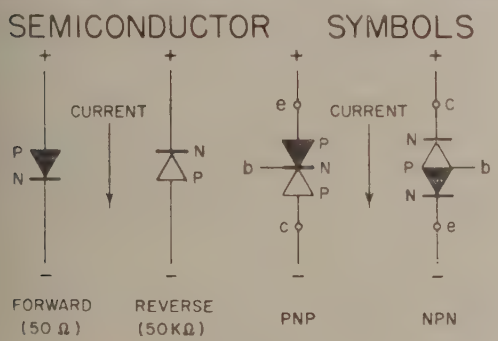


Fig. 1

On the left of fig. 1 is shown a simple crystal diode, which is called a P.N. junction in semiconductor nomenclature. If the diode is biased into forward conduction, our convention of symbols calls for the arrow to be solid. A solid arrow always indicates forward conduction. If the diode is reverse biased, then the arrow is simply outlined. An outlined arrow always shows reverse conduction.

A P-N-P transistor is shown by the third symbol. It can be seen to consist of three parts: an N type germanium body; an added P type junction which is biased into forward conduction and so has a solid arrow; and a second P type junction which is reverse biased. We call the body the base and mark it b; the PN junction in forward conduction we call the emitter, e; and the other PN junction is the collector, c.

According to semiconductor theory, a forward conducting rectifier conveys current by means of carriers of one sign alone. That is, the carriers are either electrons, or positive charges (holes), but not both. The emitter, then, injects carriers into the base. The injected carriers in a P-N-P transistor are holes. These carriers have a relatively long lifetime in the base and so most of them arrive at the collector, whose geometry is favourable for this purpose. Thus most of the injected emitter current leaves at the collector. The ratio of collector to emitter current is known as the current transfer factor α.

On the right you see an N-P-N transistor, whose symbol follows logically from the fact that the carriers are electrons, so that it has a P type base with N type junctions formed upon it. The diode symbols are therefore reversed, and the lower diode is now the emitter.

The fig. 2 shows how the well known properties of a diode can be transferred over to the transistor. At the top left are shown diodes in forward and reverse conduction. Below is shown the electrical equivalence of either, for the change in conducting state simply alters the value of R and of C. r is the so-called "spreading resistance". It is usually a constant low resistance of perhaps 50 Ω and arises from the electrical resistivity of the germanium itself.

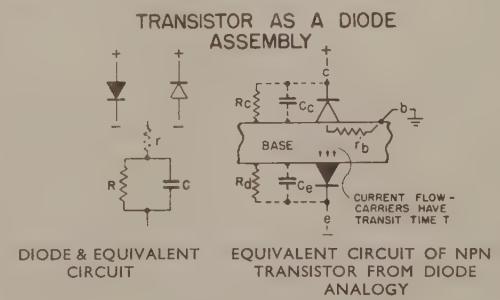


Fig. 2

On the right I have drawn a transistor semi schematically to show where the electrical components due to its collector and emitter diodes appear. The R, and the C of the collector junction are shown as R_c and C_c. When this

junction is reverse biased, as is conventional, R is high, say 100 000 Ω. C is fairly small, say 10 pF. At the emitter junction we similarly have R_d and C_e. R_d is the forward resistance of an ideal diode, often called a Shockley diode, and so its value is inversely proportional to emitter current. It is simply $R_d = 30/I_e$ where I_e is in milliamperes. R_d is 30 Ω at 1 mA.

C_e is relatively large under forward bias conditions and may be a few hundreds of pF.

The diode spreading resistance r of the two junctions is shown lumped as a resistance r'_b in the transistor base, where it appears in series with the base terminal b. Its typical value is 50 Ω.

We can use this semi schematic model to show how the transistor can produce current gain. You will recall that when we injected a current at the emitter, some of these carriers recombined before reaching the collector. Only a fraction α reached the collector. Now a transistor must obey Kirchoff's laws, so that the net algebraic current entering it must be zero. It follows that a current of $(1 - \alpha)$ of the emitter current must flow in the base lead. What is important is that if you make the base current the independent variable (instead of the emitter current), the relationship still holds: a current of magnitude $(1 - \alpha)$ caused to flow in the base will set up an emitter current of unity, and a collector current of α. There is thus a gain in current from base to collector of $\alpha/(1 - \alpha)$. This gain is commonly called β, and may have a value of 100 in modern transistors. We shall not say much about current gain in this paper. As long as it is fairly large, its exact value rarely matters in pulse and HF design work.

We shall turn now to the second major factor which controls transistor operation. This is the built-in time delay. Once we have done this, we shall have the information we need to deduce transistor properties.

We have seen that there are two classes of transistors, the N-P-N and the P-N-P. The N-P-N carries its currents by means of electrons, as in the vacuum tube. We shall therefore choose this transistor because I shall then be able to compare its delay with the delay of a vacuum tube.

Now the vacuum tube itself has a time delay, or transit time, for its carriers. If we pass an alternating signal current through the tube, and raise its frequency until the period of a cycle becomes as short as the transit time, then at such orders of frequency the characteristics start to depart from their low frequency values. In the vacuum tube the transit time is very short, say 10^{-9} to 10^{-10} s, for the electrons are free to accelerate in vacuum under a rather large anode potential. There is very little dispersion in these transit times. The high frequency disturbances do not, therefore, trouble us below a few hundreds of megacycles per second.

In an N-P-N transistor, however, the electrons must pass through a crystal lattice, where their mean free path is very small. Their drift velocity is correspondingly reduced. You will see this pictured in fig. 2, where the arrows show conventional current flow. The carriers which are electrons travel in the opposite direction. The germanium crystal through which the carriers pass is a fair conductor and is often field free. Then carriers injected at the emitter are directed to their destination, the collector, simply by their mutual repulsion. Even though the distance between emitter and collector is made as small as possible, as little as 1/1 000 in., the transit time will still be 3 to 4 orders greater than that of a vacuum tube, say 10^{-7} to 10^{-8} s, and there is a transit time dispersion of the same order. It follows that HF disturbances trouble us at a few hundreds of kilocycles per second.

In one of the most modern transistors, the drift transistor, it has been found possible to generate an accelerating field of 0.25 V or so within the crystal. By this means transit time can be reduced, so that HF disturbances are postponed until the frequency reaches a few megacycles per second.

Now how do these resistances, capacitances, and transit times influence circuit design? If you are designing an audio frequency amplifier

only the resistances, and β (the current gain) matter and you can forget the rest. We will turn to the more adventurous pulse and HF applications.

As an introduction to the pulse field it is rather easy to indicate the limitation imposed by transit time effects. Fig. 3 shows a bistable circuit which is often made the basis of a scaler.

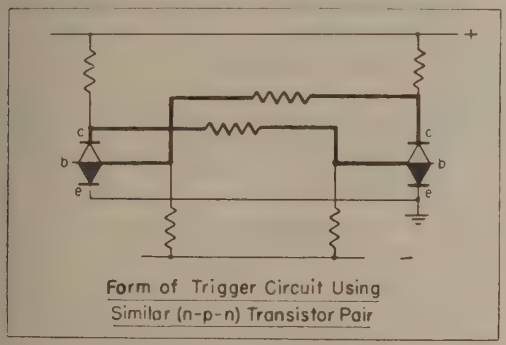


Fig. 3

As it is a trigger circuit, it must be regenerative. The regenerative loop is shown by heavy black lines. Now if you follow the loop around, you will see that it passes through two transistors. transit time delay sets a minimum resolving time for the scaler.

For the best modern transistors this time may be 2×10^{-8} s and for poor ones 2×10^{-6} s. To this time must be added delays caused by the depletion capacity of the collector, which forms a time constant with the various circuit resistances.

By careful design, however, the resolving times I have quoted can be achieved. With the best transistors 20 ns, with poor ones 2 microseconds. Of course the picture I have given you is oversimplified, but it does show how transit time can set an upper limit to speed in pulse circuits.

I shall attempt to compare the various modern transistors a little later on. In order to do this I should like to take the six parameters I have described and combine them into a few composite parameters. As you may imagine, the pulse worker will need a different yardstick from the receiver designer, and so I shall develop two sets of parameters.

On the left of fig. 4 are the parameters we have already discussed. From the first of these,

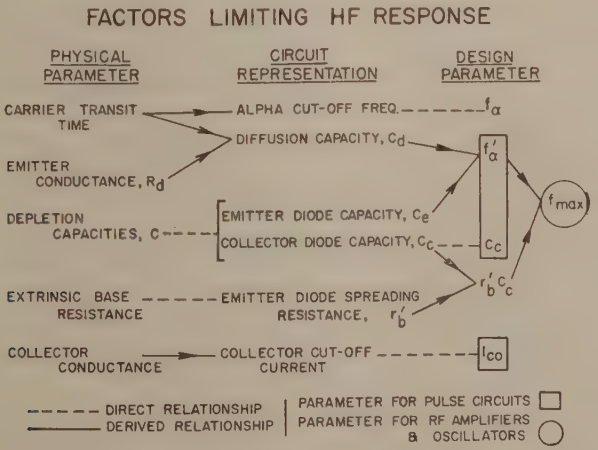


Fig. 4

Since each of these must introduce a transit time delay, a disturbance injected at one point in the loop will be delayed by at least two transit times before it returns. Evidently the carrier transit time, a frequency may be derived for which the transistor power gain will be halved. This frequency limit, f_α, is due to transit delay and dispersion alone.

In practice the gain will be halved rather earlier than this, due to the capacities present, and so modified value of f_α is used. This corrected value is derived by means of an artifice. The transit time delay may be approximately represented by a capacity-resistance combination, for such a network can also produce a time lag. Such a time constant can be placed at the emitter by shunting the emitter resistance R_d with a fictitious capacity C_d, called the diffusion capacity. C_d is so chosen that the time constant $C_d R_d$ is correctly related to the transit time effects. When we have done this, we have a means of handling the other condenser present at the emitter, that is the depletion capacity C_e. We add the diffusion and depletion capacities C_d and C_e and so form a new time constant with R_d. This new time constant yields f'_a (see rectangle). This is our first composite design parameter.

The designer of pulse circuits needs the composite parameters I have shown in rectangular boxes. In the upper box you will see f'_a, and also C_c which is simply the unmodified collector depletion capacity. The little square box at the right shows a current, I_{co}, which is a small parasitic current due to collector diode leakage. It is noted here for completeness. I shall say more about it later.

The pulse designer, then, needs f'_a, C_c and I_{co}. The receiver designer can use one single parameter as his guide. It is shown within the circle as f_{max}. It is indeed an absolute limit for him, for at the frequency f_{max}, the transistor cannot deliver more power than it receives. At this frequency the power gain is reduced to unity, and it is the limit at which a transistor oscillator will work. The receiver designer will try to work at frequencies which are not above 1/6th of f_{max}, for the transistor characteristics change too rapidly above this limit. The parameter f_{max} is quite simply derived from two time constants. One of these we have met already. It set the value of f'_α. The other is simply the product of the base resistance r'_b and the collector depletion capacity C_c.

Of the basic transistor parameters I introduced to you earlier, all but one of them are fused into this powerful criterion, f_{max}.

We have now acquired sufficient theory, and can turn to practical matters. May I just ask you to remember the composite parameters.

Here they are:

f_{max}, the frequency of unity power gain, the RF receiver designer's yardstick.

f'_α, the accurate representation of the frequency at which power gain is halved.

C_c, the collector depletion capacity.

I_{co}, a parasitic DC collector current which flows in the transistor and in the external load. It has considerable nuisance value.

2. Transistors Compared on Frequency Basis

In the next figure (fig. 5), have compared the various general types of transistor on the basis of frequency. At the top is a germanium transistor with a built-in accelerating field, the so-called graded base (or drift) transistor. You can see that it has the highest f_{max}, a few hundred megacycles/sec. Theory suggests that this limit can be raised to 3 000 Mc/s.

The pulse circuit designer sees that the f'_α is around 30 Mc/s and this is not outstandingly better than some of the other types. None of the others has an accelerating field. For transistors which are field free, f_α and f'_α are nearly identical, and furthermore, a sinusoidal signal of this frequency will be shifted by about 50° in passing through the device. Often this phase shift is quite important, and so I have shown the 50° phase shift frequency of the drift transistor, which for that type occurs at yet a separate frequency.

The second line down shows the surface barrier transistor, a device in which the germanium base is made very thin by a subtle chemical etching process. It is the nearest competitor to the graded base, but the manufacturing process does result in certain other limitations, as you will see in a later figure.

Next below you have the germanium[1] thin-base alloy transistor, whose emitter and collector diodes are fused onto the base. It is amongst the most robust and versatile of all transistors.

Then the silicon transistor which must, at the time of writing, be made by the grown process. In the grown process, the whole emitter, base, and collector structure is part of a single crystal. I shall not go into the grown process, for the limitations which will be shown in subsequent slides make the designer prefer alloyed units.

3. Feedback Amplifiers

I drew your attention to the frequency at which the transistor causes the signal to lag by 50° because I wished to say a little about the problems of negative feedback amplifier design. Such amplifiers are liable to burst into spontaneous oscillation if the phase shift is excessive. In order to avoid this danger, one design method calls for a total loop phase shift of not more than 150°, right up to the frequency where the loop gain is unity. (Feedback, 0 db.)

Now, three transistors at the 50° frequency themselves generate the limit phase shift, and the gain is still very high. Therefore attenuation must be inserted to cut the gain before this

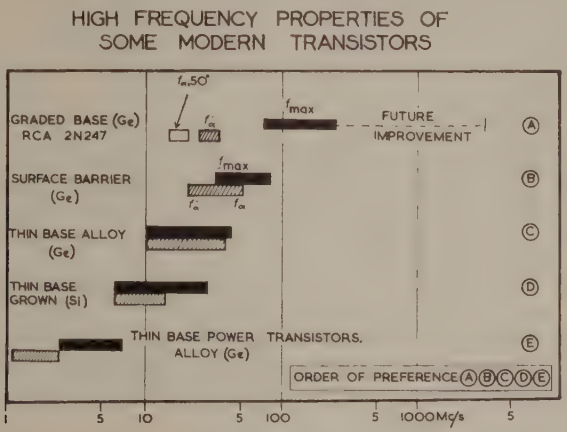


Fig. 5

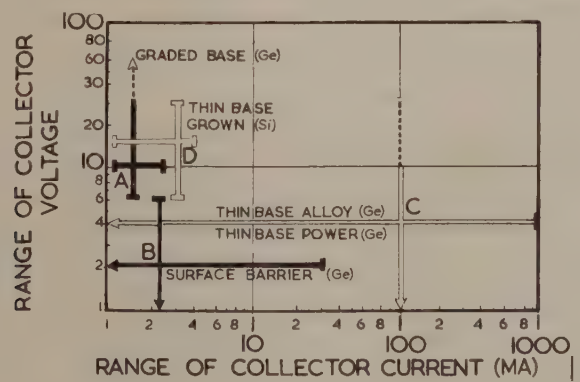


Fig. 6

Finally there is shown a thin base alloy power transistor in germanium. It is poorer than a low-power alloy unit only because of practical difficulties—theoretically it should be as good.

The transistors are here shown in order of high frequency merit A, B, C, D.

frequency is reached. Such attenuation itself must introduce further phase shift. As a result, although a non-fedback amplifier may be designed to handle 1/6 of f_{max}, it is unlikely that the pass band of a 3 stage feedback amplifier can exceed 1/10 of the 50° frequency. This sets an upper limit of a megacycle or two.

4. Transistors for Pulse Working

The pulse circuit designer will want to know what range of currents and voltages his transistors can withstand, as well as their frequency-responses.

In fig. 6, look at our alloy transistor (C) which was only third best previously. The same results also apply to the power transistor (E). The alloy type units will operate with currents from less than 1 mA to perhaps 1 A, and with voltages from a few tenths volt (or even less) up to 30 V or even a little more. (I assume you are not wanting punch through or avalanche effects to occur.)

The graded base type, previously the best, is now very restricted in scope. Its only strong point is that it can be made to withstand a high voltage. Neither it, nor the silicon unit, work well below 6 V; and neither unit will operate over a wide range of current. The surface barrier unit, on the other hand, won't stand more than 6 V on its collector before it punches through. To the pulse circuit designer, the alloy germanium transistor has the best overall characteristics.

5. Comparison: Tube and Transistor

In fig. 7 I have tried to contrast the germanium alloy transistor with a tube. You will see that whereas the power gain of a tube is often treated as infinite, the power gain of the transistor is indeed finite.

If you look at the output impedance of a transistor, you will see that it is relatively high, and comparable with a vacuum tube. However the input impedance of a transistor is very low, as you would expect, since it arises from an emitter diode and a base spreading resistance. First consider small signal conditions. If you are to realise the full power gain of the transistor, you must therefore cascade stages by means of a transformer, which matches the output impedance of one to the input impedance of the next.

At low frequencies you can do this easily. At high frequencies, matching can only be accomplished over a narrow band, since all

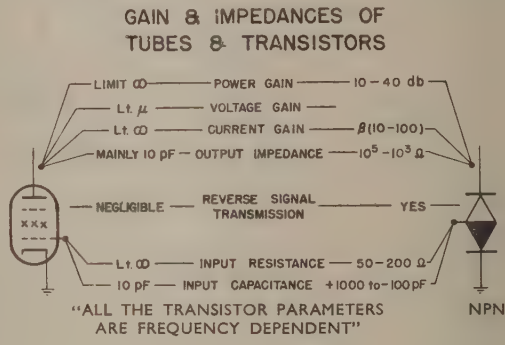


Fig. 7

transistor terminal impedances are frequency dependent. If you do try to realise the full gain at high frequencies, you will become aware of the several signal transmission which can make the reverse signal transmission which can make neutralised by external networks, which are themselves difficult to design for a wide frequency band.

6. Large Signal Wide Band (Pulse) Operation

Since matching and neutralising present difficulties over a wide band width, and are impossible under large signal conditions, the designer of pulse circuits tolerates loss of power gain in order to obtain stability and a satisfactory pulse shape. He uses another design approach. Let us first look at the approach for vacuum tubes.

The designer using vacuum tubes has only to consider voltage gain, since his grids draw no current. His key parameter is transconductance, and his voltage gain is simply the product of transconductance and anode load resistance. The latter must be chosen so that the stray capacities permit an acceptable rise time, and so the capacities set the maximum gain.

The designer using transistors will usually

employ the base as the input electrode. Thus he has a current gain β as his key parameter. He considers only current gain, since the transistor base generates little voltage in response to its driving current. If he wishes to obtain a fast rise time, he must again consider circuit capacities.

In this case the capacities must include the diffusion capacity, C_d, which we have seen can represent transit time. The capacity C_d becomes multiplied by the current gain demanded by the designer, and this is why he often cannot use the maximum value of β.

The concepts of voltage gain (for tubes) and current gain (for transistors) dominate the coupling network design in all circuits but those which involve matching. When you design a network whose current (rather than voltage) is the specified output, you may change its nature very considerably. Sometimes the tube network should be reversed for use with transistors so that the input and output terminals are interchanged.

One can recognise the experience of a transistor designer very quickly, by looking at his coupling networks!

7. Transistor as a Switch

In many pulse applications the transistor is used simply as a switch. Then its two states should be a short circuit and an open circuit. All switching devices have limitations. There is a maximum open circuit voltage which they will withstand, and often there is an associated leakage current. They rarely short circuit perfectly, and then there is a residual voltage drop, and an associated short circuit impedance.

The first line of fig. 8 compares the tube and transistor for maximum open circuit voltages, and the second line shows typical short circuit voltages. The ratio of these voltages is an important measure of the flexibility of the switch. While the open circuit resistance of the transistor can be very good, there is a parasitic current I_{co} due to leakage across the collector junction.

The saturation resistance of the transistor is seen to be far better than its tube counterpart.

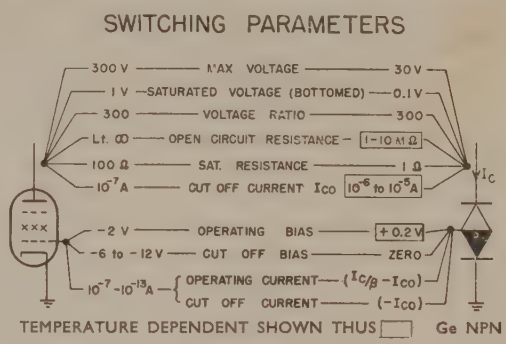


Fig. 8

The transistor is in fact an excellent switch, and it finds its greatest use in scalers, and digital computers, which are both based on two-state "on" and "off" conditions. For such work the I_{co} is troublesome, for it rises rapidly at high temperatures. The designer must always devise his circuits so that they will tolerate the I_{co} at the highest ambient temperature he will encounter.

Silicon transistors have much smaller I_{co} currents than germanium transistors, at a given temperature. The designer may have to tolerate their other limitations if the operating temperature is above 80° C.

8. Time Delay in Switches

There is a time delay in the operation of a transistor switch to which I should like to introduce you. A conducting transistor stores charge since there is a finite carrier transit time. The charge may be assumed to reside mainly in the condenser C_d of which we have already spoken. The value of C_d increases with emitter or collector current as you would expect, since the charge must also increase with current. We can therefore assess the charge in terms of the emitter and collector currents, so long as these can flow in an unimpeded manner.

Since the base current is related to the emitter collector currents by the factor β, we can also use the base current to assess charge, In fact, if the base is the control electrode, the

condenser C_d must be charged and discharged by the base current itself. While a small base current will suffice to maintain much larger collector and emitter currents in the "on" state, yet such base currents will not give a rapid change of state. Quite a large surge of base current must be injected (or extracted) to charge or discharge C_d during the transition between switched states.

When a transistor is used as a switch, the collector load is often made high so that the full collector current cannot flow. The transistor is then said to be bottomed and it exhibits the desirably low output impedance. The charge storage cannot now be based on the emitter or collector currents, but it can still be assessed, as before, in terms of the base current. The turn off time can then be computed in terms of this storage and the available surge of base current.

I have shown in a recent paper that bottomed switches need not be much slower than those using non-bottomed transistors. The reasoning is simply an extension of what I have just told you.

9. DC Amplifiers

Returning to our slide, I will draw your attention to another result of I_{co}. As 10^{-6} A in the collector it may not be too troublesome. However it finds its way out at the base, which is usually the signal electrode. This current, which is highly temperature sensitive, makes the transistor a bad device to use as a DC amplifier.

10. Auto Bias

In the second group of pulse parameters you will see the DC biases. If some sort of cathode bias resistance is necessary to stabilise a tube, can you doubt that analogous emitter bias resistance is a vital feature in a transistor, whose natural bias is so much smaller? Such an emitter resistance should develop a volt or so of bias so as to swamp the small base/emitter voltage of the transistor itself. So many circuit designers have neglected to provide this DC stabilisation. As a punishment they shall be made to use tubes!

As I have told you, the I_{co} current appears at the base. This current will traverse any impedance in the base circuit. If the impedance is a high resistance, it can generate a base to ground voltage and so disturb the bias conditions.

11. Noise

We should say a few words about noise. I_{co} is one source of noise. There is also another source. If the transistor is to operate as a useful device, there must also be a component of the standing collector current at the base. It is simply the collector current divided by β, the current gain. This component will add its component of noise algebraically to that produced by I_{co}. Rough calculations indicate that an equivalent noise source of 10 000 ion pairs would be a very good figure for a nuclear amplifier built for 1 μs pulses.

This is a large noise source indeed and it renders transistor amplifiers unsuitable as pre-amplifiers unless they follow a detector of large signal output, such as a proportional counter of fairly high gain or a photomultiplier.

12. Modulation

While the transistor does not make a good DC amplifier, yet semiconductor devices can be used to modulate small DC signals. The resulting signal can then be amplified by a transistor AC amplifier, and the DC can be recovered by rectification at the amplifier output.

A transistor switch may be used to modulate low voltage signals arising from low impedance devices such as thermocouples. In low impedance sources a small I_{co} current does not generate a significant error voltage. Such switches should be able to handle signals of 100 μV or even less.

When small currents, such as those arising in ion chambers and other high impedance devices must be amplified, use is made of a switch composed of silicon junction diodes. Silicon diodes have reverse currents whose theoretical value is around 10^{-14} A, and so should allow the modulation of currents of this order. I have published a paper on this subject, but have not attempted currents below

10^{-9} A. Present diodes are not good enough to reach 10^{-14} A, but 10^{-11} A seems a plausible target.

As I conclude this paper, I am conscious of the little I have been able to tell you, and how much I have left out. I have not even drawn an equivalent circuit. Well, I omitted the formal things deliberately, because they are in all the articles and text books.

I hope I have interested you sufficiently to read the formal approaches. When you do so I hope you will find that my approach will add something to the theoretical dry bones, and will help you to avoid some of the bypaths which would lead you away from the haven of safe, and reliable design.

DISCUSSION

KANDIAH inquired whether Moody could give some information on the behaviour of the more important transistor parameters f_α, β and I_{co} during the life-time of the transistor.

MOODY answered that he thought β and I_{co} to be the only parameters which vary appreciably. He has noticed I_{co} to vary in various sporadic ways and similar observations have been reported by the Lincoln laboratories. His practice was to design circuits for a I_{co} of $100 \div 200$ μA bearing in mind that the circuit may have to work at somewhat higher than room temperature. He feels that this practice provides sufficient margin against fluctuations in I_{co}. The factor β has been shown by many authors to be quite sensitive to surface contaminations but most of the circuits he has studied are not critically dependent on β. Although it may be that modern transistors have less surface contaminants he recommends to design the circuit conservatively so as to allow for a change of at least 2 to 1 in β.

The low frequency transistors have now proved to be so reliable that one just does not expect them to go wrong. 6 months ago high frequency transistors were not as reliable, but they were then in the early days of manufacture and there is no reason why they should not become reliable. Good results are reported with the Philco surface barrier transistor.

MOODY could not comment on the reliability of the drift transistor. As a guide to the expected life of transistors Moody quoted a failure rate of 0.1% per 1 000 hours obtained in computer work.

A discussion arose between KANDIAH and MOODY concerning the number of transistors needed on the average to replace a valve when a circuit is transistorized. Figures as high as 2 or 3 transistors per valve have been mentioned but were considered too high by MOODY and KANDIAH, who agreed that the figure was rather around 1.

In reply to a question by QUERCIA, MOODY mentioned that the noise of the transistor amplifier described in his talk corresponded to the collection of about 10 000 ion pairs. A similar figure has been arrived at by Goulding in Chalk River with a somewhat better type of transistors. For nuclear work an improvement of at least of an order of magnitude is necessary, and this is hard to get.

BALDINGER added that as far as noise is concerned the transistor behaves like a tube with a high grid current, which has high noise for high input impedance. For low input impedance a transistor amplifier can have a noise figure which compares favorably with that of a valve. He has made a 3 kc/s narrow band amplifier with a noise of 0.01 to 0.1 μV for amplifying the output of a microwave detector. It has the great advantage over a valve amplifier of comparable amplification that it is non-microphonic.

MOODY agreed that transistors are not more noisy than valves if it is possible to match the input resistance to that of the transistor. At very low frequencies, a few c/s the transistor is probably better than a valve which in this region has large flicker noise, due to spontaneous fluctuations of the cathode surface.

BALDINGER reported on measurements on the noise of transistors which show that the excess noise, behaving as $1/f$, starts to appear below about 1 kc/s whereas in a valve it is already apparent below 50 kc/s. Above these frequencies the noise is white in both cases.

KANDIAH also confirmed that the transistor is indeed better than valves as far as noise is concerned in the frequency region from 1 c/s to 1 kc/s.

MOODY referred to a recent article in Rev. Sci. Instr. on the noise of transistors below 1 c/s.

QUERCIA mentioned that he had obtained a noise of 50 μV in a 1 Mc/s amplifier.

GATTI reported on an amplifier where the gain of output voltage against charge collected at the input is stabilised by a condenser as feed back from output to input.

SCARROT inquired about the relative uncertainty of a voltage discriminator made with valves or transistors. Moody was of the opinion that the uncertainty relative to the voltage which can be applied is the same in both cases or 1/300 to 1/1 000. However for transistors it may be more suitable to use current comparison rather than voltage comparison and to apply a current bias, injected into the base or a diode, rather than a voltage bias at the grid of a valve.

KANDIAH also considered that low frequency transistors compare favorably with valves as voltage comparators

but that for high frequency transistors the allowed inverse voltage is lower, and the relative uncertainty as voltage discriminator accordingly larger

Moody pointed out that in cases where the inverse voltage is the limiting factor a diode could be introduced in the base circuit. Drift transistors have been made for up to 300 volt collector voltage and should offer good opportunities to get high discrimination accuracy.

A discussion then followed on the influence of radiations on transistors.

Kandiah reported on tests in Harwell which confirm the result of an investigation of the US Signal Corps. For doses up to about 10^{11} neutrons/cm^2, which is a reasonably high dose, there is no appreciable change in any of the parameters of most transistors. For doses of 10^{13} neutrons/cm^2 some transistors show an increase in I_{co}, a reduction in β and some signs of changes in f_a although these latter changes are marginal. The change in β may be 20% for that sort of dose. γ-ray irradiation of the intensities one might anticipate have negligible effects.

Baldinger mentioned data published in Proc. IRE July 1957 which indicate that an average transistor will tolerate 3 000 neutrons/cm^2/sec during a year before it is affected but that the sensitivity to radiation may vary by a factor of 1 000 from one semiconductor to another.

Kandiah added that fast transistors with thin bases do not seem to be affected at all.

NUCLEAR INSTRUMENTS 2 (1958) 193–202; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION V.2

TRANSISTORS IN NUCLEAR INSTRUMENTS

E. BALDINGER

Physikalisches Institut, Basel

1. Introduction

As is generally known transistors are characterized by the following advantages:

(a) small size and weight,

(b) small power consumption,

(c) insensitivity to mechanical shock,

(d) long life. The average life of lowfrequency transistors is estimated to be about 15 years, whereas for ordinary radio tubes a lifetime of about 2 years is usual.

As disadvantages one may mention:

(a) the relatively low input impedance,

(b) the difficulty of handling simultaneously high frequencies and high power.

It is evident that the transistors excel for battery operated instruments as for example portable Geiger-Müller counters, counting rate meters etc.

The advantages of the transistors appear to me to be even more important in an entirely different field of application, namely for the construction of complicated electronic devices such as pulse height analysers. As an example a 100 channel high resolution pulse height analyser constructed with radio tubes requires 1 to 2 kW power. Using transistors this value may be reduced to 100 to 200 watts. Furthermore the necessary service and maintenance of the apparatus will be greatly simplified as a result of the small heat production, the long lifetime and the insensitivity to mechanical shock of transistors. Even a 1 000 channel pulse height analyzer would still have reasonable dimensions and assured continuity of operation.

To give an example of the possible power reduction we shall consider a scale of two with a resolution of 0.5 μs.

As demonstrated in fig. 1 a reduction factor of 1 000 can be obtained. 1 000 such scalers with transistors would have a total power consumption of only 5 watt corresponding to that of a pocket flash light. In an actual apparatus the improvement by replacement of tubes with transistors will be somewhat smaller since auxillary pieces of apparatus such as mechanical counters, printers etc. must also be driven. A reduction by a factor of 10 to 50 may still be expected.

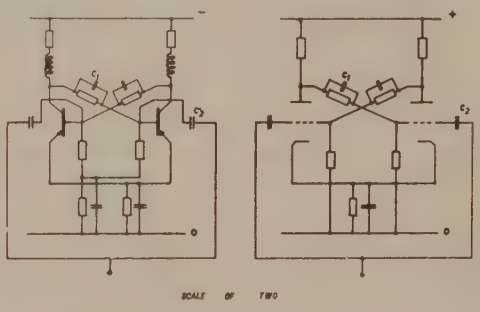


Fig. 1. Comparision of the power requirements of a scale of two with 0.2 microsec resolution (a) Tube circuit 5 watts. (b) Transistor circuit 5 × 10^{-3} watt.

2. Equivalent Circuit and some Properties of Transistors

In a triode the anode current is controlled by a voltage applied between grid and cathode. In a similar way the voltage difference between base and emitter controls the collector current in a transistor. In both cases the controlling current is only a small fraction of the controlled anode or collector current respectively. It is apparent that for both, tubes and transistors, a similar equivalent circuit should be appropriate.

The only difference are two additional resistors in the case of the transistor circuit namely:

(a) A resistor in the base lead (base spreading resistance). In the case of tubes the influence of the impedance of the grid lead can be neglected. It is easy to understand that this is no longer true in a semiconductor device like the transistor.

(b) By a similar argument we would expect a resistor between collector and base as shown in fig. 2.

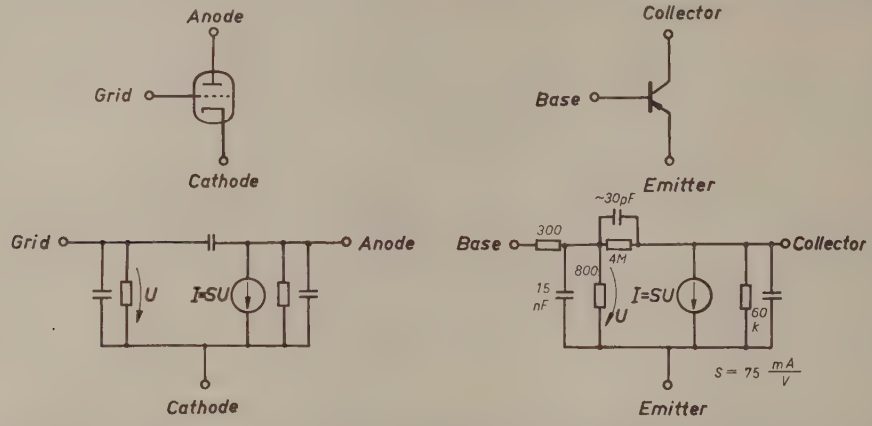


Fig. 2. Equivalent circuit of a triode and of a junction transistor for low and medium frequencies. Typical values for a low frequency transistor (type OC 71; collector current 2 mA) are indicated.

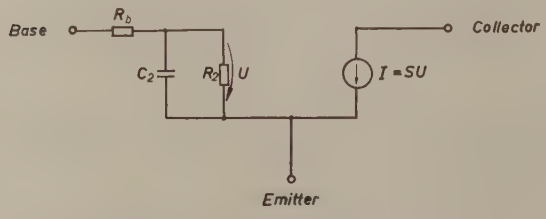


Fig. 3. Simplified equivalent circuit of a transistor.

With the assumption of a sufficiently low impedance between collector and emitter the equivalent circuit shown above may be reduced to the simpler one in fig. 3. This assumption is often justified in wide band amplifier stages and high-speed triggering circuits.

In the following section I would like to show that we have a fair theoretical understanding and a simple physical picture of the elements (small signal parameters) represented in fig. 3.

2.1. BASE SPREADING RESISTANCE

Fig. 4 shows some typical values of this resistance as a function of collector current. The increase of the resistance at small currents is readily understood. R_b can be split into two parts (compare fig. 5). The region (A) is chacacterised by a practically constant resistance. In the region (B) however the number of charge carriers increases with increasing collector current. Therefore the resistance of region (B) decreases with increasing current.

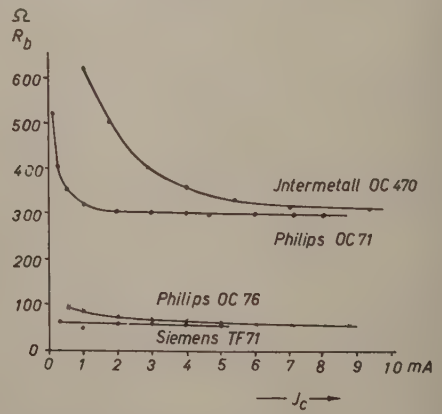


Fig. 4. Measured values of R_b as a function of collector current.

2.2. THE INPUT CAPACITY C_2

Neglecting for the moment recombination losses (only a small fraction of the holes going through the base will recombine) and assuming

a pure diffusion current, we could expect a

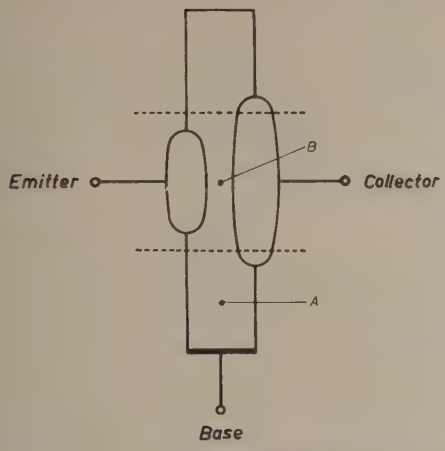


Fig. 5. Construction of a p-n-p junction transistor made by the alloy process.

linear decrease from emitter to collector of the density of holes in the base as shown in fig. 6.

For higher currents the assumption of a pure diffusion current is no longer true. An electric field is built up. Webster has shown that at higher currents also a practically linear distribution will be observed. In this case however about one half of the collector current is due to the electric field and only half is due to diffusion processes.

If we increase the collector current from I to $I + \Delta I$ a corresponding increase in the concentration of holes and electrons results and a corresponding number of + and — charges have to be added in the base region. Evidently the base region can accumulate charges in a manner similar to a parallel plate condenser. In our case the electrodes of the condenser are given by the conduction and by the valence band of the semiconductor material. The input capacity is obtained by dividing ΔQ (Fig. 6) by the corresponding variation of the emitter-

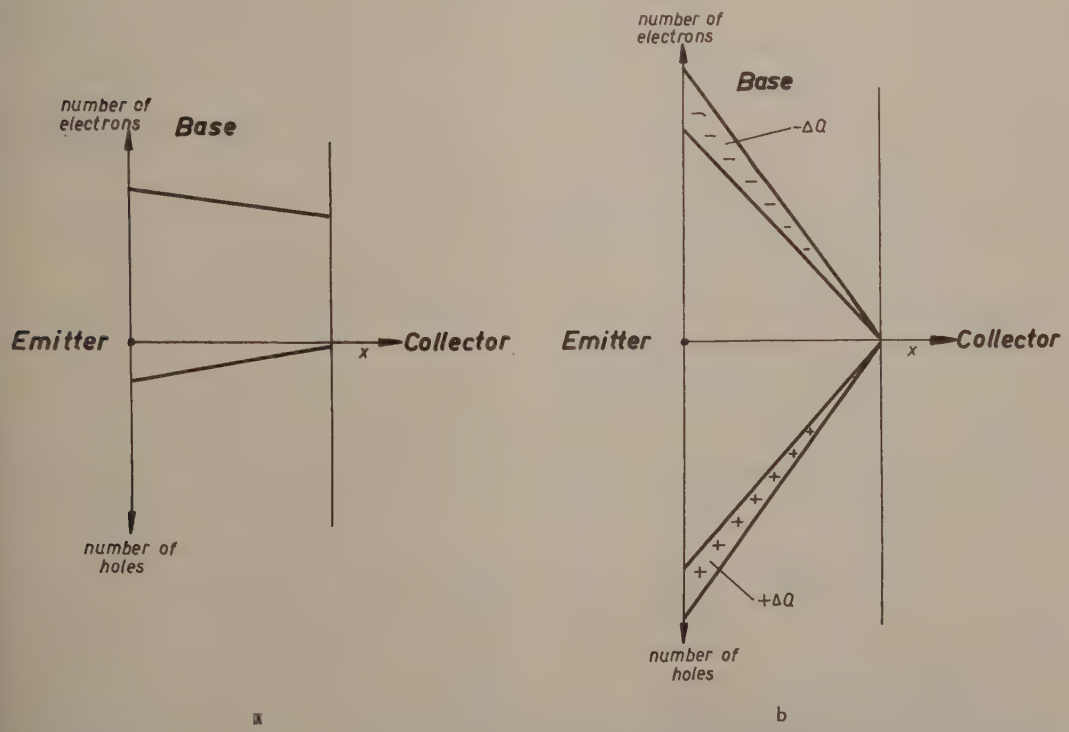


Fig. 6. Number of holes and electrons in the base region of a p-n-p transistor as a function of x (only schematical). (a) for small current. (b) for high current.

junction voltage†. From this picture we immediately expect a linear increase of the input capacity with increasing current and that the slope is twice as great at higher currents compared to the slope at lower currents. At higher currents only half the current is due to diffusion and only the diffusion current will give a contribution to the input capacity.

This typical behaviour is nicely confirmed by fig. 7.

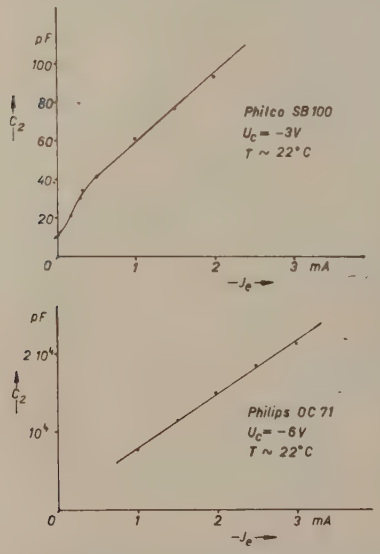


Fig. 7. Input capacity as a function of collector current for a low frequency and a high frequency transistor.

2.3. THE TRANSCONDUCTANCE S

The transistor may be said, to consist of two diodes; the emitter diode biased in the forward direction and the collector diode biased in the reverse direction.

Shockley has shown, that for a junction diode the following equation gives a good description of the diode characteristic:

$$I = I_s \left(e^{\frac{q \cdot U}{k \cdot T}} - 1\right) \qquad (1)$$

where:

q = elementary charge
k = Boltzmann constant
T = absolute temperature
U = voltage applied to the junction
I_s = saturation current in reverse direction.

For higher currents in the forward direction the 1 in the bracket can be neglected. The above equation may be applied to the emitter junction in a transistor and we expect therefore a transconductance S of

$$S = \frac{\partial I_c}{\partial U_{ej}} \cong \frac{q}{k \cdot T} \cdot I_c \qquad (2)$$

U_{ej} = voltage applied to the emitter junction.

The transconduction of a transistor is to a good approximation independent of type, manufacturer and material and will have the same value for low frequency and for high frequency transistors, for Ge- and for Si-transistors. These facts are demonstrated to be true to a good approximation by the measured values shown in fig. 8.

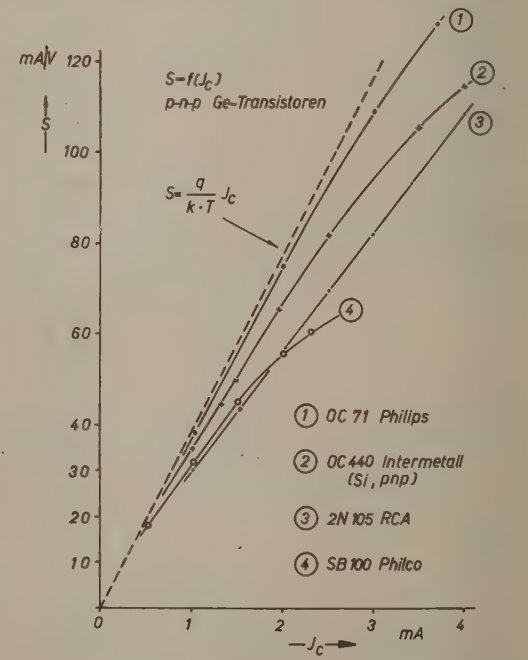


Fig. 8. Transconductance S for some transistors as a function of collector current.

I would like to mention that we may occasionally observe rather large deviations from this simple behaviour (compare fig. 9). It may be shown experimentally that a slight modification of the eq. (2):

† Note that in C_2 also the depletion layer capacity is included. This contribution is generally small and nearly constant in the region shown in fig. 7.

$$S = \frac{q}{k \cdot T} \cdot m \cdot I_c \qquad \frac{1}{2} \leq m \leq 1 \qquad (3)$$

and the assumption of an emitter lead resistance R_e ($R_e \neq f(I_c)$) gives in most cases a reasonable agreement between experiment and theory.

If the measured values shown in fig. 9a are plotted as shown in fig. 9b it is possible to

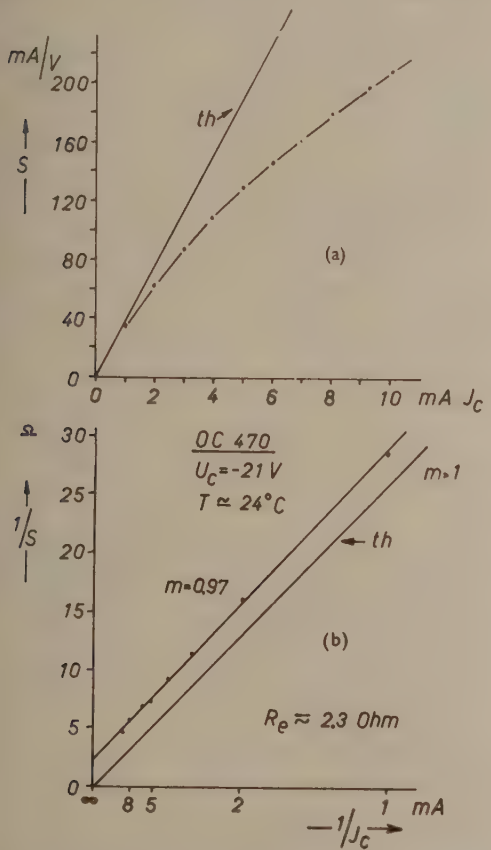


Fig. 9. (a) Measured values of S (definition of S compare eq. (2) and fig. 3) as a function of collector current. (b) $1/S = f(1/I_c)$. If the measured values are lying on a straight line, the emitter lead resistance and the factor m in eq. (3) can be determined.

determine m in formula (3) and R_e. The measured points in fig. 9b lie on a straight line and thus confirm the two assumptions mentioned above.

2.4. THE RESISTANCE R_2 AND THE CURRENT AMPLIFICATION FACTOR $\beta = S \cdot R_2$

The resistance R_2 in fig. 3 allows for the driving current necessary in the base lead. Several components contribute to the resultant base current. In a p-n-p transistor these are: (a) electrons flowing from the collector to the base, (b) electrons flowing from the base to the emitter, (c) electrons disappearing by volume and by surface recombination, (d) electron-hole pairs generated by thermal agitation.

It is thus apparent that the base current is rather complicated and it will not be discussed further. The conception of the product $\beta = \Delta I_c/\Delta I_b = S \cdot R_2 =$ current amplification factor is much simpler and typical values are presented

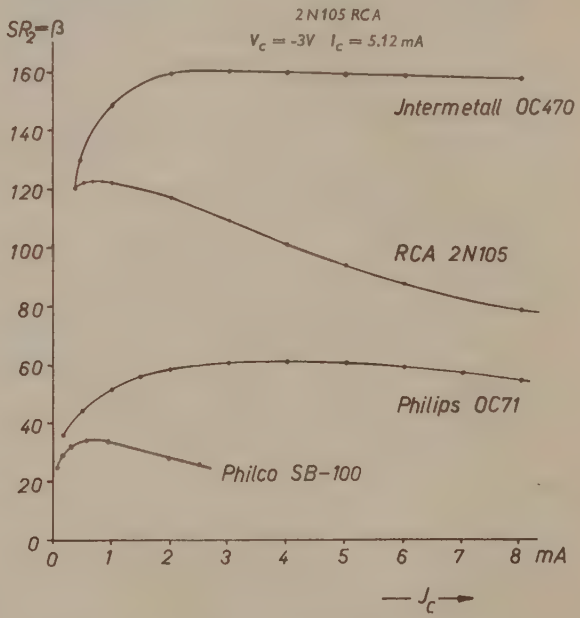


Fig. 10. Typical values of the small signal current amplification factor $\beta = \Delta I_{\text{collector}}/\Delta I_{\text{base}}$.

in fig. 10. The smaller values of β at higher and at lower currents are due to the decreasing emitter efficiency and to surface recombination respectively.

3. Temperature Dependence of Transistor Parameters

The influence of temperature on the small signal parameters of the equivalent circuit is rather weak and under normal laboratory conditions will produce no special difficulties (compare figs. 11 and 12).

Most difficulties observed in practice are connected with a unwanted DC-shift of the operating point. For these reason the variation of the DC-emitter-junction voltage U_{ej} and of the DC-base-current with temperature have to be considered carefully. Typical curves are presented in figs. 13 and 14.

These figures demonstrate that to each collector current corresponds a temperature T_1 where $I_b = 0$, and a temperature T_2 where $U_{ej} = 0$. At higher temperatures both quantities change their sign. With this behavior in mind it is relatively simple to anticipate the DC-shift of the operating points in transistor circuits.

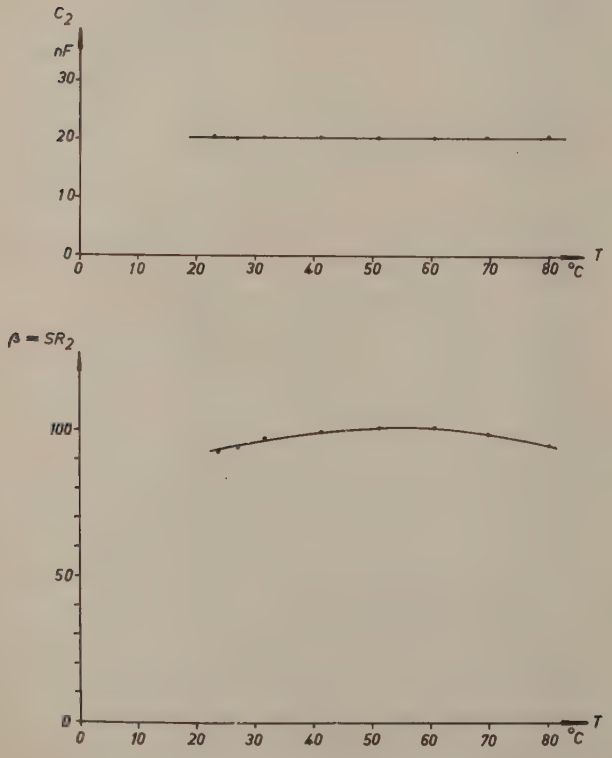


Fig. 11. Input capacity and current amplification factor β as a function of the junction temperature T. [2 N 105], $V_c = -3V$, $I_c = 5.12\,mA$.

4. Some Simple Circuits

4.1. TRANSISTORIZED THERMOSTAT

The temperature variation of the junction voltage and the base current may be used to construct a thermostat (fig. 15).

Transistor no. 1 operates simultaneously as a temperature sensitive element and as an amplifier, driving the power transistor no. 2. All circuit elements are located inside the thermostat. We believe, that such a thermostat may be useful for DC-amplifiers and for other transistorized precision measuring circuits.

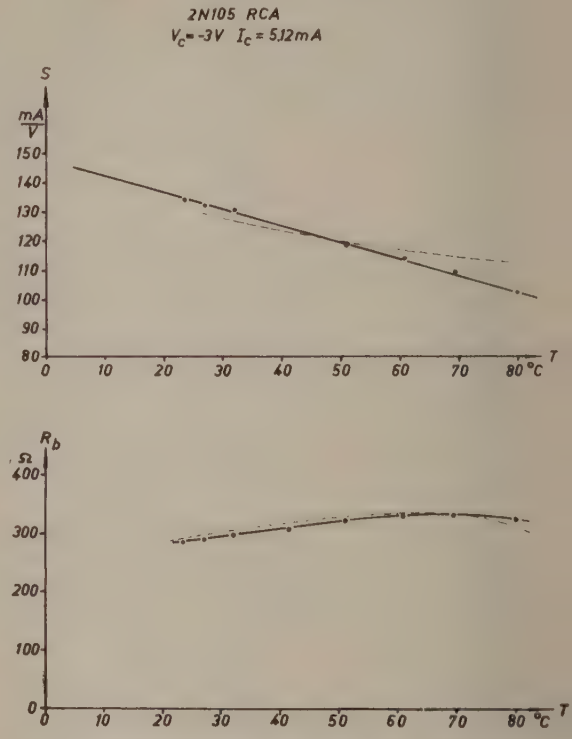


Fig. 12. Transconductance S and base lead resistance R_b as a function of the junction temperature. Note that eq. (2) predicts $S \propto 1/T$ (the broken line in the fig. for S indicates $1/T$ dependance) whereas the measured values show a stronger influence of temperature.

4.2. GATE CIRCUIT USING NO SUPPLY VOLTAGE AND ONLY ONE TRANSISTOR

A positive input pulse applied to the emitter will be transmitted to the collector. A positive gate signal applied to the base will cut off the transmission.

4.3. SCHMITT-TRIGGER CIRCUIT

The well known Schmitt-trigger circuit is often used as a precision amplitude discriminator. With ordinary vacuum tubes a long term

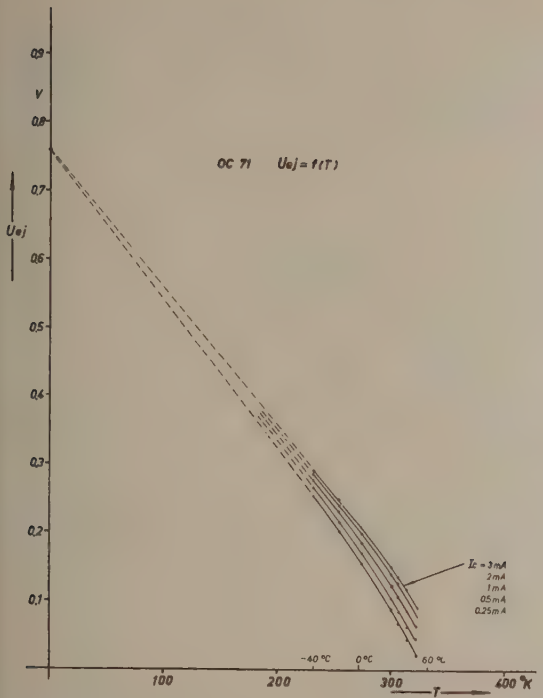


Fig. 13. DC emitter junction voltage U_{ej} as a function of the junction temperature with collector current I_c as a parameter. Note that all extrapolated curves cross the point $T = 0$ at $U_{ej} \cong 0.76$ volt. The value of 0.76 volts corresponds to the energy gap of germanium.

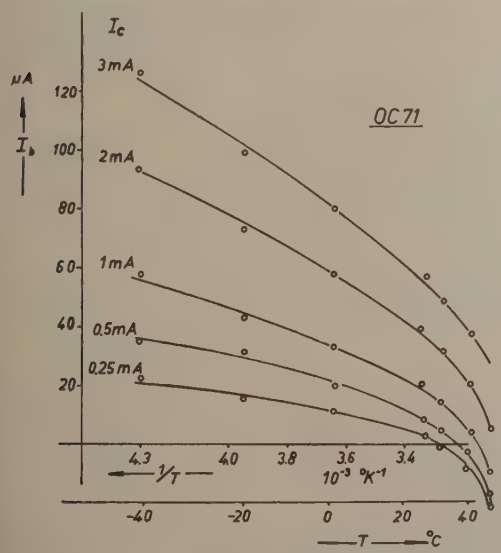


Fig. 14. DC-base current as a function of temperature with collector current I_c as a parameter.

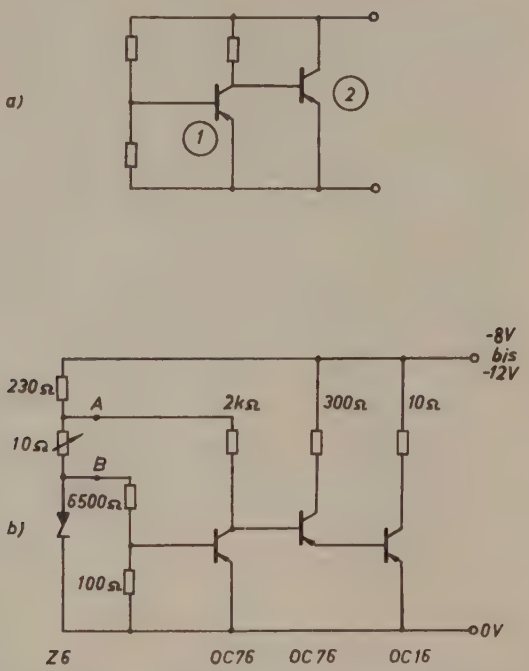


Fig. 15. Thermostat using transistors. (a) Basic circuit. (b) Practical circuit using a Zener diode for the compensation of the influence of supply voltage fluctuations. Without compensation the regulated temperature depends on the value of the supply voltage.

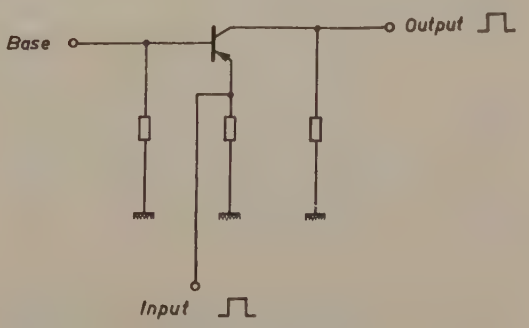
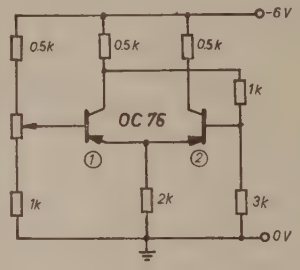


Fig. 16. Gate circuit.

drift of the triggering levels of about 0.1 volt is observed. Assuming an input pulse height of about 100 volts, a precision of 1 part in 1 000 is thus possible. We may ask: how good is the precision of as transistorized circuit of the same type? Fig. 17 demonstrates the variation with ambient temperature of the triggering levels of such a circuit measured with a DC-input voltage.

It is essential that the temperatures of both transistors are the same. A temperature difference of 10°C produces a corresponding shift



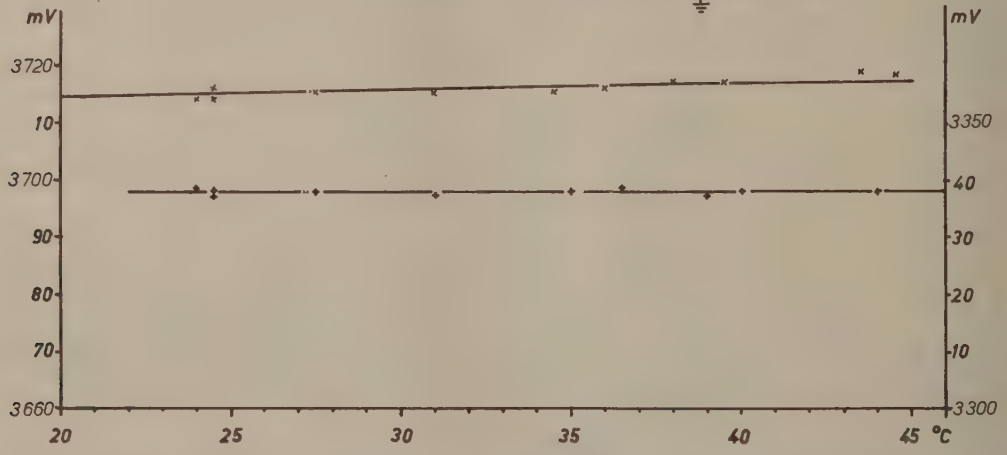


Fig. 17. The two triggering levels of a transistorized Schmitt-circuit as functions of ambient temperature measured with DC-input voltage. The ordinate to the left shows the voltage between base and ground of transistor no. 1 for the higher and the ordinate to the right for the lower triggering level.

in the triggering levels of about 20 mV. If pulses of short duration are used as an input signal, the junction temperatures, and consequently the triggering levels, will therefore depend to some extent on the pulse repetition rate. For the circuit shown in fig. 17 this influence is estimated to be of the order of $\frac{1}{3}$ mV. These measurements demonstrate that a transistorized Schmitt trigger circuit permits at least the same precision as the corresponding vacuum tube arrangement.

4.4. POWER SUPPLIES

With transistors available at present stabilized voltage supplies may be designed which display highly satisfactory characteristics, e.g. very low output impedance, high regulation factor of several thousands and small temperature dependance. As in tube circuits two types of stabilizing networks are possible.

As reference voltage a Zener-diode may be used. Fig. 19 shows a circuit of the shunt type, characterized by the following data:

Output voltage variable:	1–16 volt
Max. load current:	2.5 A at 1 volt
	1.7 A at 10 volt
	1 A at 14 volt
Stabilizing factor $\eta = \frac{\Delta U_E}{U_E} \cdot \frac{U_A}{\Delta U_A}$	$\approx$ 1 000
Output impedance	$\approx$ 0.001 Ω
Ripple voltage	$\approx$ 1 mV
Temperature coefficient of the output voltage	$\approx$ 1 mV per 1°C

The reference voltage is derived from a small stabilized power supply (indicated in fig. 19 by a broken line) which uses a Zener diode and additional resistors to compensate for line voltage fluctuations. It may be mentioned that the power supply shown in fig. 19 can withstand a sustained short circuit of the output terminals.

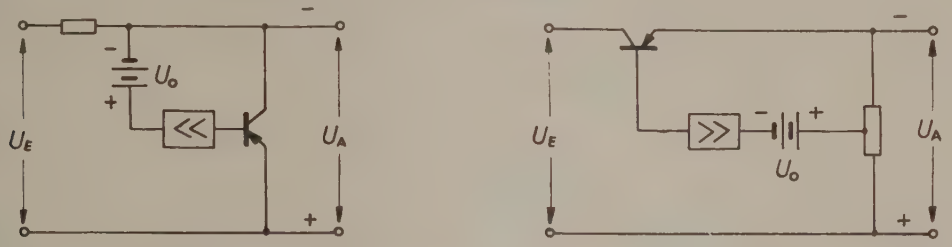


Fig. 18. The two types of stabilized voltage supplies. (a) Shunt connection. (b) Series connection.

For literature see for example:
A. W. Lo *et al.*, Transistor Electronics (Prentice-Hall, 1955); Transistors I (Published by RCA Lab. Princeton N. J., 1956); R. D. Middlebrook, An Introduction to Junction Transistor Theory (John Wiley & Sons, 1957).

The curves shown are results of partly unpublished measurements made in Abteilung für angewandte Physik, Physikalische Anstalt der Universität Basel. Most of this work has been supported by "Stiftung Hasler–Werke", Bern.

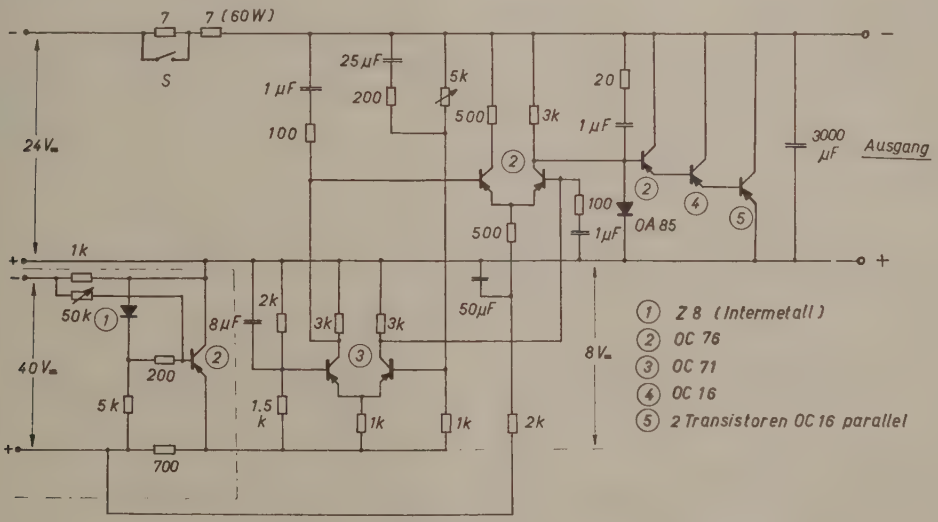


Fig. 19. Circuit diagram of a stabilized voltage supply of the shunt type.

DISCUSSION

MOODY remarked that Baldinger's equivalent circuit is certainly not incompatible with the picture he gave in his own lecture. The input condenser of Baldinger's model is directly connected with the transit time, since the charge stored in the base is equal to the current flowing through the base multiplied by the transit time to go across the base. The cut-off frequencies are directly connected with the average value of the transit time and its fluctuations.

BALDINGER agreed that the transit time effects are lumped in the input time constant of his model and remarked that usually the dispersion in the transit time is short compared to the input time constant, and can be lumped together with it without the need for an additional *RC*-circuit or delay line—except for silicon grown junction transistors, in which the transit time dispersion cannot be neglected.

MOODY pointed out the great advantage of the Baldinger model that the parameters he introduces are frequency independent. On the other hand these parameters refer to the internal structure of the transistor whereas the more conventional parameters are directly connected with what appears at the terminals.

MOODY pointed out that Baldinger had used the voltage characteristics of the transistor.

BALDINGER agreed that in many cases it was more fruitful to consider current because of the low input impedance.

MOODY raised the point of the temperature and time dependence of I_{co}, which he said is often very peculiar and may cause instability in a system.

GATTI pointed out that I_{co} instability in trigger circuits can be overcome by feeding the supply voltage to the base through chokes so that there is a balancing of voltage between the two base-emitter junctions. Moody considered this a very useful procedure.

BALDINGER considered that base current instability is a less serious problem than one might think since it is always possible to find an operating point where the base current is quite naturally defined by emitter efficiency.

MOODY pointed out that the condition for zero base current is that the I_{co} is compensated by the collector current represented by β, so that the small base current comes out as the difference of two large currents both of which are relatively unstable. The laboratories of Bell Telephone have reported that small amounts of water vapor in the envelope can cause drastic changes in β and I_{co}.

BALDINGER felt that these instabilities are certainly very serious in circuits involving large numbers of transistors which all have to be highly reliable, but that in individual measuring instruments where only a few of the transistors are used in critical applications one can afford to select the transistors for best stability.

KANDIAH confirmed that good time-stability can be obtained with transistors working under zero base current conditions in a thermostat on similar lines as that described by Baldinger, on the condition that the critical transistor is selected out of a batch for absence of surface contamination effects.

On a question BALDINGER explained that the parameter I_{co} does not appear in his analog model since it is a small parameter and is taken into account in the DC base current. This DC base current is the sum of several components. In a pnp transistor we have the electrons going from the collector to the base, and from base to emitter, holes and electrons recombining by surface or volume recombination, and electrons and holes created by thermal agitation. I_{co} therefore has a rather complicated origin.

MOODY showed samples of two plug-in transistor units. These are designed for a computer being constructed in Canada, which will be entirely composed of units of these two types, as follows.

1. A pnpn trigger circuit, which is a combination of one pnp and one npn trigger circuit. The time required for switching it on is between 0.05 and 0.1 μs. It can deliver any load up to 50 mA. At present it is used for switching at intervals of 1.5 μs, but operation at shorter interval is possible.

2. A diode logic gate circuit.

MOODY also showed a sample of a miniature transistor servo amplifier. This consists of two parts, an aluminium base on which the power transistors are mounted, and a printed wiring board. After connection, the space between the parts is filled with plastic. The circuit ensures that the power dissipation in the germanium transistors is only $\frac{1}{4}$ watt at the full output load of 3 watt, and the unit will operate at a temperature of 85–95° C.

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SESSION V.3

A TRANSISTORISED MEMORY LINE

D. MAEDER

Eidgenossische Technische Hochschule, Zürich

Since we were chiefly interested in the construction of circuits rather than in the manufacture and internal secrets of tubes and transistors, we tried to look at transistors in the transistor trigger circuits, but these did not compare well with the performance of equivalent tube circuits. After that, the application of transistors was limited to a proportional pulse

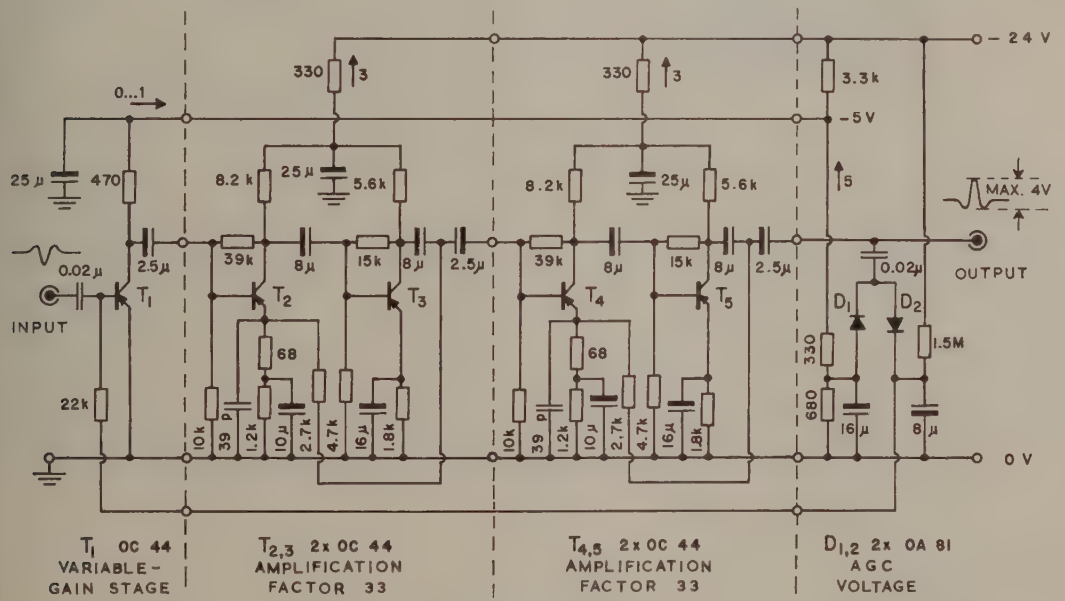

Fig. 1. Transistor Pulse Amplifier with AGC.

same way that we look at a list of tubes selecting the type suited for a specified purpose. However, we have tried, in our nuclear instrumentation, to restrict the application of transistors to such circuits where certain well-known advantages of transistors make their use so tempting that there could be no hesitation in employing them. These advantages have been recapitulated to-day by Prof. Baldinger in his paper.

In 1955, a few early trials were made with amplifier in the micro-second range at low signal levels.

Our amplifiers are used in memory signal receivers where the input signal is usually only a few milli-volts. Since the amplifier is required to accept pulses about one micro-second, long, having rise-times of a few tenths of a micro-second, it was necessary to use the fastest transistors at present available at a reasonable price, namely the OC44.

The amplifier, shown in fig. 1, is constructed

from two units each containing two type OC44 transistors, with a conventional feed-back network (2.7 kΩ, 68 Ω) which limits the low-frequency voltage gain of each unit to 33. The

often used in vacuum tube circuits. The effect of the combined feedback networks is to reduce to a negligible value the dependence of amplification factor upon temperature.

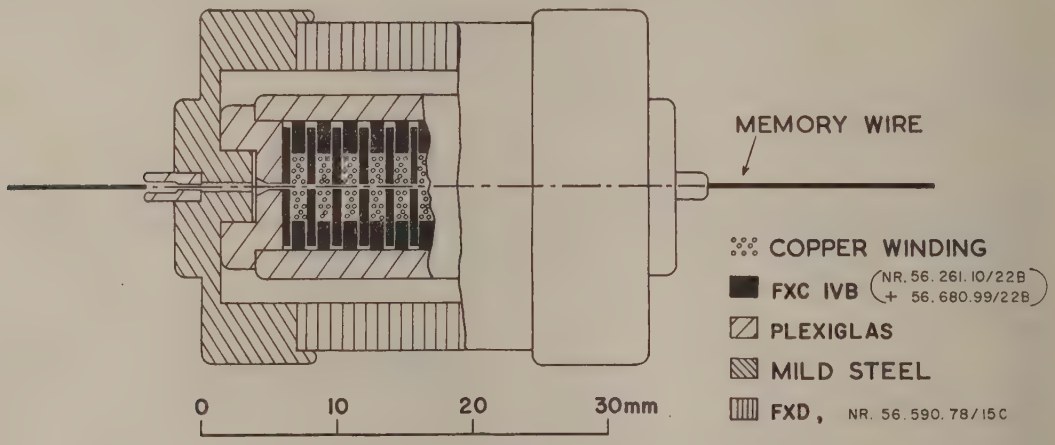

Fig. 2. Section through electro-acoustical transducer

emitter by-pass capacitor value must be adjusted to give acceptable high-frequency performance (with some early batches of OC44 a value of 390 pF was necessary, but with more recent

The input and output impedances of the amplifier are defined as the input series impedance and output parallel impedance which halve the output signal but preserve its shape.

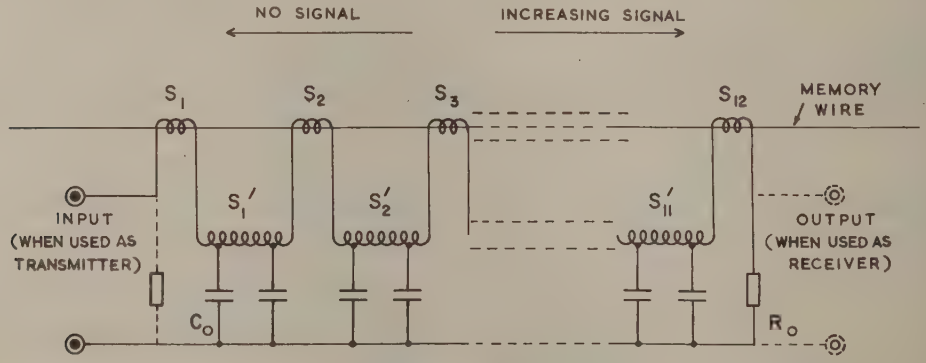

Fig. 3. Interconnection of the coils.

ones a smaller value between 10 pF and 50 pF is necessary to give good pulse shapes). In our applications, we usually tolerate a 5% overshoot when an ideally flat step-function from a very low impedance source is applied to the input. The quiescent-state current in each transistor is stabilised by a network (39 kΩ, 10 kΩ), similar to a grid bias network which we have

The input impedance was found to be equivalent to 8 kΩ in parallel with 50 pF. This impedance is, of course, much higher than the input impedance of a transistor without feedback, where the effective shunt capacitance would be several hundred pF. The output impedance was found to be equivalent to 40 μH in series with 40 Ω.

Since the amplifiers are used for low signal

level applications, considerations of noise are important. It was found that about 20% of the type OC44 transistors had to be rejected because of excessive noise, but the remainder were satisfactory.

Two such amplifier units have been combined with an input transistor stage in which the bias is regulated from the final output, giving automatic gain control. This complete assembly is then used as the input receiver for the signals from a magnetostrictive memory line, using transducers of the type shown in figs. 2 and 3.

Longitudinal mechanical waves travelling along a steel wire are received by the transducer which contains 12 coils. If the coils are phased properly, an electric wave is built up in one direction. The amplitude of the wave will increase each time it passes through one of the coils just as in a distributed amplifier. Thus one obtains an increase of the signal and the phasing conditions lead to a substantial reduction of unwanted reflections. Since this is true at both the input side and the output side, noise problems are radically eliminated and one can even use materials which otherwise would have too small a magnetostinctive effect. We have found that a stainless steel wire still provides enough signal. We can make additional improvements by interchanging the positions of some of the coils, or by mismatching at the end of the chain, so that the slight dispersion which is still present may be counteracted to some extent. Thus, we have obtained a pulse shape which should permit storage of 2 000 bits on a memory line of about 3.5 ms length.

Fig. 4 shows the shape of the output pulse after amplification in the transitor amplifier. With an input in the order of 10 mV, the output pulse was maintained by the amplifier AVC at a constant level of approximately 4 V peak to peak.

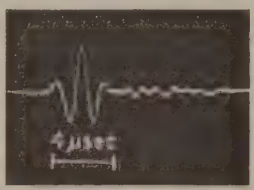

Fig. 4. Pulse received from an 18 m steel wire memory.

The distance between the two negative peaks is about 1.8 μs, and this determines the closet spacing at which digits can be placed in the line. Thus for this line of 3.5 ms total delay, there is a digit-capacity of about 2 000 digits.

The same amplifier units have also been used in an electrostatic memory system. Although the input impedance is somewhat low for this purpose, a satisfactory signal pick-up from a Williams electrostatic memory tube has been obtained.

DISCUSSION

Scarrot: If you make a coil which has a high Q, which will make a good delay line, then it is a very bad transducer, and if you make a coil which is a good transducer, it has to have a low Q and has therefore got no energy left to handle the next coil. How do you design individual coils?

Maeder: On fig. 2 you can see that we have very little air around the coils. The black areas are of ferrite material. The only air gap is between the wire and the hole in the ferrite discs. The hole diameter is about 0.3 mm, whereas the wire has 0.2 mm ϕ, leaving an airgap of only 0.05 mm. If we only used the transducer coils to make up the delay line, we should need too much delay per section to match to the acoustical delay, and the result would be a very low Q delay line. Fig. 3 shows that we actually introduce between every two of the transducer coils an auxiliary coil to construct a high Q delay line, and thus we obtain a reasonably high impedance (e.g. 500 ohm) with the correct delay. One can even introduce some inductive coupling by means of the taps on the auxiliary coils in order to improve the phase characteristics of the complete delay line.

Maeder replied to a question of Dr. Quercia that the noise has not been measured. Input signals were normally about 10 mV, and not less than 1 mV. At these levels the noise was insignificant, and he thought it was below 100 μV, but was unable to give further details.

Quercia stated that he made some measurements with a pulse amplifier of this type. He found that the transistor type OC44 was not the best for such purposes, since it had a very high α value of about 0.99, and consequently its cut-off frequency is lowered. He preferred to use a transistor which he believed to be Sylvania type 2N135, which has a lower cut-off frequency, but also a lower value of α. With

this combination of properties the maximum frequency of the amplifier might be made higher than that for a value proportional linear amplifier.

MAEDER replied that this might be important, if one wished to amplify very small signals, but fortunately he had not had that problem.

SCARROTT asked whether the transmitted signal was from transistors, or from a valve source.

MAEDER replied that they were no longer using nickel, but now used steel, in order to have less dispersion. However the efficiency of conversion was at least ten times smaller for steel than for nickel, and consequently greater input power was required which could not be obtained with transistors. A very inexpensive television pentode had been used.

HOOGENBOOM said that this difficulty could be avoided by using steel wire plated with nickel, and this could be bought.

MAEDER replied that he had tried this technique, but that the nickel stripped off during the process of winding coils.

REAVES asked whether Dr. Maeder had also built an equivalent vacuum tube amplifier, and if so, how did the cost and performance of the two compare.

MAEDER replied that the great advantage of the transistor amplifier was that it could be supplied from a battery or from an extremely well filtered power supply, to avoid the introduction of interference. With vacuum tubes, interference of the order of millivolts might be introduced, for example through the heater, and this was one of the main reasons for eliminating vacuum tubes in that application. Transistors were used here not because they were available, but because they were really much more convenient.

KANDIAH asked what current was required in the transducer coils for driving the wire.

MAEDER replied that there had in fact been several different coil arrangements, one of which was of 200 ohm impedance, and another of 500 ohm. The 200 ohm version was usually preferred, and this had only one auxiliary coil between each pair of receiving coils. However, a better quality delay line was obtained by the use of more auxiliary coils, and this also permitted the insertion of receiving coils at unorthodox positions to correct for unwanted peaks. Therefore the 500 ohm version, was generally used and gave the pulse shape shown in fig. 3. The current supplied to the 200 ohm version in 0.3 μs pulses was 250 mA, and this was transformed down in the ratio 2 : 1 by a ferrite transformer. Thus a signal of about 100 mV was obtained from a delay line consisting of nickel wire 0.2 mm dia. and 5 m length, although this was not of good pulse shape. With a very long line no signal could be obtained, because the attenuation of the nickel then used was quite appreciable. From a line of steel wire of 0.15 mm dia. and 5 m length a signal of 40 mV was obtained, but it was thought that the quality of steel could influence performance. Therefore for a very short length of wire, nickel is preferable for this reason, but for a long length of wire, steel is definitely preferable to nickel on account of its lower attenuation.

A discussion followed between MAEDER and SCARROTT on the attenuation in steel and nickel wires. SCARROTT reported some figures he obtained in a very similar problem. The inefficiency of the transducers accounted for a factor 1 000 between the voltage levels at transmitter and receiver. In addition there was the attenuation in the wire to be taken into account amounting to a factor 10 for a 4 ms torsional wave delay line. Maeder reported that he had measured an attenuations of 100 in a nickel line but only a factor 10 in steel. Scarrott recalled that the torsional waves have 60% the speed of longitudinal waves and therefore need a shorter wire for a given delay. The damping is probably smaller than for longitudinal waves.

VON DARDEL asked how the wire was mounted, and whether difficulties had been found with its suspension.

MAEDER replied that since the adoption of this delay-line arrangement he had had no difficulties of this type, since the delay line produces a travelling wave in one direction only. Waves in the reverse direction are mainly cancelled, so it is relatively unimportant if the end is left open, or clamped, or damped. There must, of course, be some damping.

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SESSION VI.1

STANDARDIZATION IN THE FIELD OF NUCLEONIC INSTRUMENTS

H. I. PIZER

CERN, *Geneva*

The title of this session may need some clarification, so I will commence with a short resumé of the origin of this subject as it has arisen in CERN.

You will realise that CERN being an international organization, purchases equipment from most of the 12 Member States and also from across the Atlantic, and in addition we build our own equipment. CERN will also have visiting teams of physicists coming to do experiments on the Cyclotron and the Proton Synchrotron, and they will in general bring their equipment with them. The purchased equipment and the equipment developed in CERN must be compatible—by this I mean that the input and output voltages, polarities, impedances must be such that where required the equipment can be interconnected in the desired way in order to carry out experiments.

In the case of purchased equipment we have already found it necessary to make modifications, to alter connectors and, in some cases, impedances: this of course occurs in all laboratories. In the case of equipment brought to CERN by visiting physicists they will normally be self-contained, but we must expect that they will need to borrow equipment from our laboratories either to replace faulty equipment or in order to modify their experiments.

From CERN's point of view, some sort of standardization in the field of nucleonic equipment would be of great service in reducing the number of modifications which we have to make and increasing the interchangeability of the nucleonic equipment; this is especially important for experimental equipment with the Proton Synchrotron which will be operating in 1960. We think that if this standardization could be achieved it would save in physical and intellectual effort on the part of all subsequently concerned with experiments.

Similar problems will arise around other high energy machines now under construction such as in Harwell, Saclay, Rome which will certainly attract research teams from other laboratories. The creation in Great Britain of the National Institute for Research in Nuclear Science raises similar problems. Of course, one can argue that these latter problems only point towards national standardization, and I think you will be hearing from later speakers the plans of standardizing equipment in particular countries, but it would certainly be a useful step if it could be made an international one.

It was with this idea in front of us at CERN that a series of meetings were held which commenced in January 1956. The participation was restricted simply because it was thought that it would be very difficult to reach any sort of agreement in the early stages if all opinions were represented. These meetings were attended by representatives from CERN, from the French Nuclear Research Centre at Saclay, from the British Atomic Research Establishment at Harwell, and from the University of Liverpool. The objects of these meetings were to promote co-operation and to assist standardization of nuclear equipment for use with particle accelerators. The minutes of these previous meetings are available for any who would like to read them.

I would like now to give you an account of the results of these meetings. At that time in 1956 both Harwell and Saclay were considering independently a programme of development of a new range of nucleonic instruments and so it

was a very convenient time in which to discuss the details of the instruments and the co-operation for standardization.

The problem of whether these two series of instruments were mechanically similar was investigated and it was soon clear that the post-office rack system which is so widely used in both Europe and America was in fact an international standard, and that no major difficulties would arise on this account. For the details—Saclay were using a plug-in sub-chassis system; at that time Harwell were not using such a system, but I believe that a system of sub-chassis will be used in the future though the dimensions will be somewhat different.

By comparing the actual instruments which were being developed by these two establishments, it became clear that broadly they were proceeding along similar lines and that in general the individual instruments were being developed to carry out similar functions.

I have drawn up a table of nucleonic instruments which are being used in CERN, Harwell and Saclay, or which are being developed for future use. This has already been circulated but extra copies are still available, and I now have details from other countries and institutes, and I hope to issue a more complete table including this information at a later date.

However, in order to make these boxes completely interchangeable, the inputs and outputs must be identical both electrically and mechanically. This implies identical interconnecting plugs and sockets and similar electrical impedances where coaxial cables are used, and the polarity, amplitude and shape of the signals must be similar.

Some discussion took place on the possibility of arriving at some common coaxial connectors, but no actual agreement was reached on this point. The difficulty in arriving at a standard for connectors is both a financial one and a national one. Whereas in CERN we are able to employ well tried American connectors, both Great Britain and France are forced generally to use a locally produced article. In some cases the American connector is made under licence, but then one is faced with a further difficulty which is of a national character, since it often occurs that one is obliged to satisfy some national standard requirements, particularly since at least some of this equipment will either be used by the Armed Services, or simply because they must work within the confines of Armed Services general specifications.

At these meetings we did attempt to adopt the same connectors as were used by the North Atlantic Treaty Organization but this proved impossible since at that time no such specifications were in existence. Eventually we were forced to accept that either the connectors would have to be changed where necessary, or that special interconnecting cables would be used when it was necessary to mix the equipments.

An important decision was taken on the impedances which should be used for pulse work with coaxial connections. There were two values chosen—125 ohms, and 75 ohms. The decision of 125 ohms was taken because 200 ohms was going out of favour due to the faster rise times which are encountered with the new photomultipliers and associated equipment. 125 ohms was in use at the University of California Radiation Laboratory at Berkeley, and both medium and high quality cables were available from the U.S.A. at this impedance. It was generally agreed that a lower impedance should also be chosen as a standard and that 75 ohms was suitable.

It is not an easy matter to arrive at an agreement on any sort of standard, and in the case of cable impedances we were probably rather hasty in our decision, because since this agreement two of the participants have found valid reasons for preferring 100 ohms to the 125 ohms.

On the subject of standardization achieved there is not a great deal further to relate. On the subject of co-operation I think we felt rather more satisfied. The meetings led to an exchange of ideas and some measure of agreement on the division of development effort to

avoid duplication. Since the meetings have been held alternately in each of the participating institutes they have given the members good insight into the facilities and activities of these respective institutions.

The experience of these meetings between Harwell, Saclay, Liverpool and CERN, and the interest shown by other European institutes indicated that the time had come to have discussions on a wider basis.

By these previous meetings we have set the stage for the possibility of a wider participation on standardization. We in CERN are particularly keen to hear the views of representatives of CERN member states, and also of our visitors from overseas. I think it is clear that at this session either the subject of a European standardization will gather momentum because you feel that it is a worthwhile subject to put effort into, or it will die because perhaps the problems of arriving at suitable standards are almost insuperable.

We in CERN would certainly like to hear the opinions of the member states, and I would ask you to join in the discussions so that we can get a feeling of your opinions.

NUCLEAR INSTRUMENTS 2 (1958) 210-211; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION VI.2

A GENERAL SURVEY OF ELECTRONICS IN THE FIELD OF NUCLEAR PHYSICS

I. F. QUERCIA

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Previous speakers have described many elegant electronic devices developed by specialists in electronics. But it is possible to view the application of electronics in a different way, from the point of view of the physicist who makes use of the devices in the course of his experiments. This survey is only a tentative sketch of such an outlook, and doubtless others may wish to make additions to it.

The physicist finds that the use of detectors of high energy nuclear events usually involves the use of electronic devices. Whilst no new information is here presented, it is useful to group the known types of detectors in categories as shown in the following table.

Group p (pulse)	Group t (track)	Group i (integrating)
Type		
1. Scintillation counter	1. Nuclear emulsion	1. Ionization chamber
(a) transmission		
(b) integral		
2. Čerenkov counter	2. Bubble chamber	
3. Proportional counter	3. Diffusion chamber	
4. Geiger counter	4. Wilson chamber	
5. Fast ionisation chamber		

It may be profitable to review the present state of the art of electronics as applied to each of these types of detector. From the viewpoint of the experimental physicist in the high energy field, we may then analyse for which types there exists a weakness requiring further development of the electronic techniques.

For detectors of group p, it can be said that they use "inherent electronics", i.e. that the use of electronic devices is unavoidable, and inherent in their detection techniques. With type 1 and type 2 a transducer is used in the form of a photo-multiplier. With the remaining three types, the inherent electronic circuits are coupled directly, without the aid of a transducer.

It is here suggested that the associated electronic circuits available are adequate for employment with the detectors of group p now in use, but that the weakest part of these devices lies in the design of the photomultiplier itself. The possibility of collaboration between CERN and a commercial organization, for the purpose of developing improved European photo-multipliers, would be worth discussion.

Detectors of group t are associated with a type of inherent electronic system, which for types 2, 3 and 4 takes the form of interlock and control circuits. These circuits are now very well developed, and it may conservatively be claimed that they are adequate for use with these detectors.

But there is another type of electronic system, which we may label "auxiliary electronics", and which might be developed for the purpose of providing a greater speed of scanning of the

information obtained from detectors of group *t*. In this connection, automatic electronic devices might be used either for the pre-scanning of a large number of events in order to select those requiring subsequent measurement, or for the detailed measurement of the small number of selected events. For example, with a Wilson chamber ten thousand photographs might be obtained, of which only ten photographs contained tracks worthy of detailed measurement for a particular experiment. It is here suggested that for such a case it is more important to develop the former type of automatic pre-scanning electronic device for selection, than to develop the latter type of measurement device.

For detectors of group *i*, which have some type of inherent electronic circuit associated, it could probably be claimed that existing electronic circuits were adequate for the purpose.

DISCUSSION

Moody stated that many adaptors between different types of coaxial connector were in use in his laboratory in Canada, and a Mr. Bradley could give information about them.

De Waard supported the suggestion that more development on photo-multipliers should be carried out in Europe. While the new RCA type 6810A was attractive, a further need for basic development still remained, and CERN might play a part.

Quercia suggested that a tube manufacturer might regularly visit CERN to study the requirements of physicists. For example, Prof. Bernardini had proposed that a photo-cathode window might be made of a glass of which the refractive index matched that of a Čerenkov solid glass counter, to minimise losses on contact.

Surdin said that in France the firm La Radiotechnique also were studying the production of fast photo-multipliers. Their present types were similar to RCA and Dumont types, and next year they expect to make one similar to the RCA 6810 with a curved outer surface for the photocathode.

De Waard regretted that in France there had hitherto been no novel development of photo-multipliers, but only a repetition of known types of American design.

Surdin explained that commercial demand and import difficulties conditioned the manufacturing programme in France.

Meunier said that he personally had used the photo-multipliers of La Radiotechnique, and had found them faster than the RCA tube, with one additional stage, and with ability to work at higher voltages. Thus there was less "spread" inside the tube, and it was not just equivalent to the RCA tube.

De Waard asked whether this French tube was then as good the new RCA tube.

Surdin replied that at present it was not, but that next year it would be better. French tubes now exist with 14 stages, but with a smaller output current than the RCA tube. However, next year they will give an output current of 300–400 mA, proportional, and they will have a curved cathode.

Gatti suggested that a useful new type of photo-multiplier would be one which combined early stages of dynode type with later stages of venetian blind type, for when the number of electrons is high, statistical spread is of less importance. The following amplifier would not be a very fast one, and theoretical analysis shows that this does not worsen the time resolution.

Rispoli said that the earlier papers presented by Gatti and Meunier had shown that for use with scintillation counters and Čerenkov counters it was possible to construct electronic equipment of very high resolving power, of the order of 0.1 ns. However, both the detector and the photo-multiplier were inferior to this, and therefore there was now greater need to develop the detector methods and devices than the electronic circuits. Much work was needed on the development of detectors and photo-multipliers, and it might be desirable to consider the possibilities of quite new methods and materials, and new types of faster photo-multipliers.

Åström said that it would be desirable to develop improved photo-multipliers which were not only faster but also more sensitive, particularly for use in spectroscopy. At present an input of about 10 photons is necessary to give an output of 1 photo-electron. It would be of great value if this sensitivity could be improved by a factor of 10.

Kandiah said that past experience made him doubt whether such specialised development work could be commissioned as suggested. Only large industrial laboratories could undertake this work, and the economics of such a proposition were unattractive. Since under the more favourable conditions in America no conclusive photo-multiplier design had yet been reached, it was not safe to assume that we could make any more rapid progress in their development.

Von Dardel thought that national establishments such as Saclay and Harwell were better placed than CERN to press their manufacturers for development effort. However, in contrast, the work started in Zürich by Schaetti and continued by Baumgartner demonstrated the possibility of successful development in a small laboratory.

NUCLEAR INSTRUMENTS 2 (1958) 212–215; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION VI.3

LA STANDARDISATION DES APPAREILS ÉLECTRONIQUES AU COMMISSARIAT A L'ÉNERGIE ATOMIQUE

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1. Introduction

L'utilisation d'appareils de physique nucléaire constitués d'éléments fonctionnels standardisés présente d'indéniables qualités parmi lesquelles ressortent une grande commodité d'exploitation et des avantages économiques nombreux, résultant à la fois de la standardisation et de la division des ensembles en petites unités.

Le Commissariat à l'Energie Atomique possède dans ce domaine, une expérience de près de 10 ans, qui a abouti en 1954, à l'adoption d'un système de mécanique convenant à ses besoins. Notre propos est de décrire ce système et d'examiner les réalisations qu'il a permises. Nous étudierons successivement:

Les différents éléments standardisés et leur constitution (§ 2).

La façon dont ont été localisées les fonctions élémentaires et les combinaisons qui résultent de leur association (§ 3).

Les efforts qui ont été accomplis pour assurer une haute sécurité de fonctionnement (§ 4).

2. Standardisation de la Mécanique

La cellule élémentaire est le tiroir 1 unité (fig. 1), d'où sont dérivés les tiroirs 2 et 3 unités. Des tiroirs prolongateurs, des trois dimensions, permettent d'avoir accès au câblage d'un tiroir particulier aux fins de mise au point ou de dépannage (fig. 2). Le bâti standard peut contenir 5 tiroirs unitaires et est conforme aux normes américaines, (hauteur 5 unités de $1\frac{3}{4}$ inch ou 44.45 mm). Enfin les baies sont constituées d'éléments de 10 unités pouvant être accouplés dans le sens vertical et le sens horizontal. La fig. 3 montre une baie de 4 éléments.

Ce matériel normalisé a été conçu avec le souci de préserver au maximum l'espace disponible, de permettre une grande liberté dans la disposition du câblage, de nécessiter le minimum d'usinage et d'utiliser largement les profilés normalisés de faible prix de revient. Sa réalisation a été publiée en 1955 d'une façon détaillée[1]). Depuis cette époque, le jeu d'élements disponibles a été complété et amélioré par la création d'un montage pivotant à glissière télescopique, de panneaux à ventilateurs, d'un pupitre amovible et tout récemment, d'éléments plats de hauteur égale à 2 unités, en diverses profondeurs, dont les platines de câblage verticales permettent d'assurer une excellente ventilation des lampes (fig. 4). Enfin, il existe des baies monoblocs de 40 unités destinées aux installations à poste fixe et qui sont, la plupart du temps, scellées au sol.

3. Éléments Fonctionnels Standardisés

Parmi la multitude de réalisations basées sur l'emploi de tiroirs standardisés certaines sont utilisées très couramment et sont considérées comme éléments fonctionnels standardisés.

Ces éléments sont caractérisés par:

(a) La normalisation des tensions d'alimentation et du brochage des connecteurs.

(b) Une étude minutieuse des circuits, souvent complétée par les enseignements de l'utilisation normale.

(c) Une grande sécurité de fonctionnement (point sur lequel nous reviendrons à la dernière partie).

[1]) R. Fabre et R. Viguié, l'Onde Électrique XXXV, 344 (1955) 1108–1115.

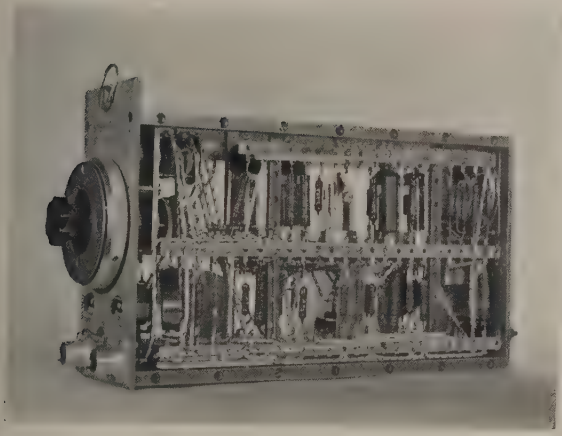

Fig. 1

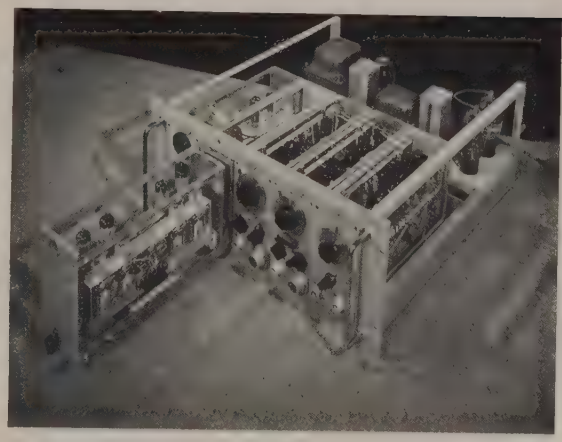

Fig. 2

Fig. 3.

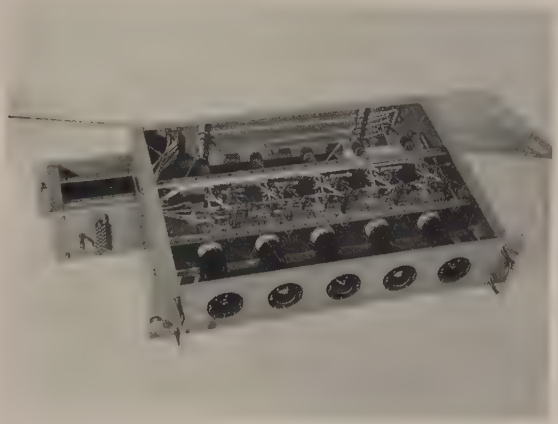

Fig. 4

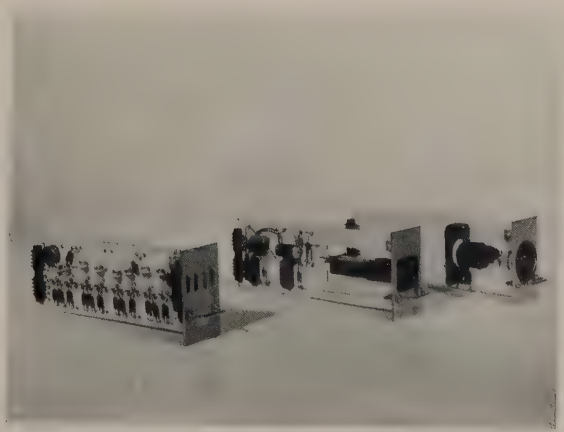

Fig. 5

Fig. 1. Exemple de tiroir untaire (tiroir intégrateur).

Fig. 2. Montage à tiroirs interchangeables, avec tiroir prolongateur, portant un tiroir double decade.

Fig. 3. Une baie de quatre éléments.

Fig. 4. Nouvel élément plat.

Fig. 5. Exemples de tiroirs transistorisés.

(d) La normalisation des signaux d'entrée et de sortie, dans la mesure du possible.

Chaque tiroir standard accomplit une ou plusieurs fonctions élémentaires choisies parmi les plus courantes, de telle sorte que la majeure partie des besoins est couverte[2]). L'association de ces éléments fonctionnels permet de réaliser une multitude de combinaisons, la plupart du temps sans prendre de précautions spéciales concernant les alimentations et les interconnexions. L'étude de ces combinaisons est facilitée en remarquant que le "tiroir universel" (comportant les fonctions sélection d'amplitudes, mise en forme, coïncidence ou anticoïncidence) peut constituer le point d'articulation de toute installation de Physique Nucléaire. En amont de cet élément se trouvent le détecteur et la chaîne préparant convenablement les informations délivrées par celui-ci; en aval, sont disposées les voies de dépouillement et de présentation des résultats.

L'emploi de tels éléments fonctionnels présente une grande utilité non seulement pour le physicien, mais également pour l'ingénieur chargé d'étudier un appareil complexe.

4. Securité de Fonctionnement [3])

Souvent pour l'Energie Atomique et, en particulier, pour la physique nucléaire, ce n'est pas tellement le prix d'un appareil qui importe le plus, mais sa robustesse et sa durée de vie. Le problème revêt, pour des raisons de sécurité ou d'économie d'exploitation un caractère souvent impératif. De l'avis de nombreux utilisateurs d'appareils, physiciens ou ingénieurs, une grande sécurité de fonctionnement est plus importante que des performances sensationnelles sauf, bien entendu, dans certains cas spéciaux où les objectifs sont généralement plus limités.

L'ingéniosité essentielle dans l'étude ou le projet d'un ensemble électronique consiste dans l'appréciation exacte des limites de performances des éléments utilisés.

Il faut aussi savoir choisir et introduire avantageusement les différentes fonctions élémentaires, leur nombre étant toujours réduit au minimum.

Enfin l'attaque d'un circuit par des signaux à polarités positives ou négatives peut avoir pour effet, sous certaines conditions, de diminuer les problèmes de dissipation.

Les techniques courantes permettant d'assurer dans le domaine des appareile électroniques utilisés en physique nucléaire une parfaite sécurité de fonctionnement, ne diffèrent pas essentiellement des principes utilisés dans les autres domaines de l'électronique (télécommunications, radar, machines à calculer).

L'ingénieur chargé de l'étude et surtout celui chargé de la fabrication industrielle doivent choisir une bonne pièce détachée qui devra toujours être utilisée avec des coefficients de sécurité suffisants, les caractéristiques limites de fonctionnement ne devant jamais être atteintes.

L'accessibilité et la rapidité de remplacement sont essentielles étant donné les fréquences actuelles des pannes. C'est l'une des raisons pour lesquelles, le Service des Constructions Electriques du C.E.A. a standardisé le matériel C.E.A sous forme d'éléments ou tiroirs fonctionnels. L'interchangeabilité ainsi obtenue s'est avérée avantageuse et indispensable pour les utilisateurs et les services d'entretien mais elle n'est naturellement pas une condition essentielle pour la sécurité de fonctionnement.

4.1. REGLES GENERALES

Afin d'illuster les quelques règles générales énoncées ci-dessous nous donnerons rapidement par la suite quelques exemples (pièces détachées, circuits, disposition de câblage) de réalisations des ingénieurs d'études ou de fabrication ayant essentiellement pour objet l'amélioration de la sécurité de fonctionnement.

Du point de vue construction on peut recommander généralement:

(1) Utilisation des circuits les plus simples possibles avec de grandes marges de tolérances.

(2) Choix d'une pièce détachée de haute qualité.

(3) Ne pas faire travailler les divers éléments

[2]) H. Guillon, l'Onde Électrique XXXVII, **361** (1957) 388–391.

[3]) Communiqué par M. Guyot C., S.E.I., S.C.E., C.E.N. Saclay.

(condensateurs, résistances, tubes électroniques etc.) aux limites de leurs caractéristiques.

(4) Réduction des consommations au minimum.

(5) Utilisation des techniques modernes (par exemple tubes électroniques développés en fonction d'une utilisation spéciale, transistorisation, circuits imprimés, etc.).

(6) Remplacement des éléments actifs par des éléments passifs (exemple les tores à ferrites pour remplacer des mémoires électroniques, lignes à retard, etc.).

(7) Ventilation toujours convenable des armoires et des baies.

(8) Stabilisation des tensions alternatives.

(9) Effectuer des essais sérieux sur les prototypes industriels ou de laboratoire.

(10) Remplacer ce que l'expérience a révélé comme défectueux par des systèmes ou des principes nouveaux possédant ou garantissant une plus grande sécurité de fonctionnement.

(11) Dans des ensembles complexes l'expérience prouve que l'incorporation de systèmes permettant des vérifications marginales est nécessaire. Au cours des visites d'entretien périodiques ces systèmes permettent de déceler rapidement les éléments ayant évolué d'une façon trop considérable. Leur remplacement systématique permet d'éviter des interruptions au cours de l'exploitation, l'interruption pouvant se prolonger plus ou moins longtemps.

Du point de vue exploitation on peut dire que:

(1) Le fonctionnement en continu des appareils est recommandable.

(2) Un entretien périodique systématique, ou entretien préventif (par la méthode des tests marginaux) permet d'assurer, pendant un intervalle de temps d'utilisation donné, une sécurité de fonctionnement plus grande.

(3) Les éléments remplacés devront être de bonne qualité, sélectionnés et préalablement vieillis.

4.2. EXEMPLES PRIS SUR LE MATERIEL NOUVEAU STANDARD C.E.A.

Les exemples qui suivent montreront quelques points ou le souci de la sécurité de fonctionnement a conduit à adopter telle ou telle solution lors de la construction du matériel dit "nouveau standard C.E.A."

Pièces détachées

La pièce détachée utilisée est généralement conforme aux normes U.T.E. (Union Technique Electricité) et C.C.T.U. (Comité de Coordination des Télécommunications de l'Union Française). Les tubes électroniques sont du type "longue durée" des séries "Noval" ou "Miniature". On a utilisé des tubes développés en vue d'applications spéciales ou utilisant des techniques modernes:

emploi de diodes au silicium à haut isolement type THP 8 ou 14 P2; tubes Dékatron type GC 10/BS et type GC 10/4B; enfin, on peut signaler l'utilisation prochaine des transistors dans les tiroirs numérateurs.

Disposition de câblage

A ce point de vue, le principe de câblage et la disposition des éléments du tiroir double décade (fig. 2) ont été adoptés en vue de pouvoir utiliser très prochainement la technique des circuits imprimés.

Pour améliorer la résistance d'isolement du circuit d'intégration dans le tiroir intégrateur TIS-1 certaines connexions ont été réalisées systématiquement en fil nu, rigide de diamètre 12/10 mm, de façon que les câbleurs ne soient pas tentés de faire ramper ce file long du châssis. Les cosses relais sur lesquelles sont soudées ces connexions sont encadrées par des conducteurs portés à un potentiel voisin de celui du circuit d'intégration en vue de jouer le rôle d'anneau de garde. On améliore ainsi notablement la sécurité de fonctionnement en éliminant tous les effets possibles des variations de résistances de fuite.

Conception des circuits

En ce qui concerne la conception des circuits un effort a été fait en vue de déterminer les schémas offrant la meilleure sécurité de fonctionnement possible; on se bornera à citer deux exemples:

(a) L'expérience a montré qu'il est préférable que la sortie de la pompe à diodes du tiroir intégrateur s'effectue sur une anode, plutôt que sur une cathode en raison des variations de

l'isolement entre filament et cathode des diodes.

(b) Dans le tiroir double décade on doit signaler le principe de réaction adopté entre la quatrième et la première bascule: cette réaction se fait par un montage utilisant deux diodes au germanium type OA 85. Ce montage permet en effet, des variations importantes des caractéristiques des cristaux et la décade fonctionne correctement pour des valeurs des résistances inverses de ces diodes descendant jusqu'à 150 000 Ω.

4.3. TRANSISTORISATION

Pour améliorer la sécurité de fonctionnement du matériel C.E.A., le Service des Constructions Electriques s'est efforcé aussi d'utiliser les transistors; ces éléments ont d'abord été adoptés sur le matériel le plus urgent (appareils portatifs et appareils de prospection), ensuite, sur le matériel utilisé en très grand nombre (échelle de 1 000, tiroir numérateur) (fig. 5).

Du point de vue sécurité de fonctionnement, on a pu constater que les appareils à transistors offraient incontestablement une résistance plus grande aux essais de vibrations. D'autre part, les transistors étant soudés directement sur le circuit il n'y a pas de pannes dues à de mauvais constacts (support de lampes, broches etc.); il est, en outre, possible de protéger l'ensemble du câblage par un revêtement convenable qui assure une meilleure résistance aux agents atmosphériques (humidité, vapeurs acides etc.). Utilisés dans les conditions normales de température, de gros espoirs sont fondés sur les nouveaux appareils C.E.A. à transistors (tiroirs numérateurs, échelle de 1 000, sélecteurs multicanaux).

4.4. FREQUENCE RELATIVE MOYENNE DES PANNES

Les fréquences moyennes relatives des pannes par an par appareil sont indiquées sur les tableaux ci-joints (tableaux 1 et 2).

Les chiffres figurant sur le tableau 1 sont relatifs aux appareils C.E.A. utilisés en régime de fonctionnement continu sur des installations fixes. Le tableau 2 donne les chiffres pour les appareils utilisés en laboratoire de façon discontinue. Pour les types d'appareils choisis la fréquence des pannes est sensiblement 35 à 40% plus élevée dans le tableau 2; ces résultats sont relatifs à 750 appareils réparés.

Tableau 1

Frequence moyenne relative des pannes

(Appareils en fonctionnement continu)

Type	Panne/an/appareil (1956)
Echelle de 1 000	2.5
Intégrateur échelle	3.1
Amplificateur 2 Mc/s	2.3
Amplificateur à courant continu	1.7
Tiroir entrée	0.55
Tiroir décade	0.35
Tiroir intégrateur	0.8
Tiroir numérateur	0.4
Tiroir T.H.T.	0.35

Tableau 2

Utilisation non continue en laboratoire

Type	Panne/an/appareil (1956)
Echelle de 1000	3.5
Intégrateur-échelle	4.1
Amplificateur 2 Mc/s	3.1
Amplificateur à courant continu	2.3
Tiroir d'entrée	0.75
Tiroir décade	0.5
Tiroir intégrateur	1.0

Les tubes électroniques restent toujours les éléments les plus vulnérables. Sur une installation fixe, comptant plus de 30 appareils et représentant environ 600 tubes électroniques (type longue durée ou sécurité), 67 pannes ont été enregistrées durant 7 557 heures de fonctionnement par suite de la défaillance des tubes et 91 pannes durant 13 830 heures, représentant respectivement 51% et 43% les causes de pannes. Après 14 000 h de fonctionnement tous les tubes électroniques des appareils installés dans ce tableau fixe ont été changés systématiquement.

Dans les installations non fixes des laboratoires 80% des pannes enregistrées sont dues aux seuls tubes électroniques.

NUCLEAR INSTRUMENTS 2 (1958) 216–217; NORTH-HOLLAND PUBLISHING CO. – AMSTERDAM

SESSION VI.4

UNITISED EQUIPMENT AT A.E.R.E.

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Although various types of unitised equipment have been in use in many laboratories there has not been an attempt to introduce a rigid system of mechanical unitisation in Europe until Saclay proved a sound system of this type. The developments at Harwell owe much to the convincing demonstration of this system at Saclay. There were however some features of the Saclay system which, we felt, could be improved upon. The major improvement was in the provision of a standard method of ventilation of the individual sub-units which thermally isolated each horizontal layer in a rack from the units above or below it. To this end a tray unit was developed which acted as a support for the individual sub-units and also as a method of cooling the individual layers by convection.

Views of a high level pulse amplifier, a non-overloading pulse amplifier using double delay line shaping, a single channel pulse analyser and a distributed amplifier are shown in figs. 1, 2, 3 and 4. It is worth noting the following major differences between these and the Saclay units. Firstly the units are much deeper and occupy the full length of the short depth rack. Secondly the individual sub-units are slightly wider than 3″ so that 5 such sub-units occupy the full width between the vertical members of a standard rack. Thirdly the heater supplies to individual units are derived from heater transformers of the correct ratings mounted on these units. Some of these differences accrue from the knowledge that the heater current required in individual units is variable from 0 up to possibly 3A or 4A. It is not possible to cope with these variations from a heater transformer without an unduly large variation in heater voltage. In the interests of reliability it is necessary to maintain the heater voltages in valves within 4% of the rated voltage—this being the maximum permissible tolerance in heater voltage on long life valves.

All metal sheet parts are vertical in the interest of good cooling thus providing a smooth flow of convection current. Thermionic valves are mounted horizontally in this convection stream.

Fig. 5. shows the standard tray unit. We have decided to use a common mains switch on this tray unit with two outlets to the plugs. One line is energised as soon as the switch is turned on and the other is delayed by approximately one minute by means of a delay unit which is fitted inside the tray unit. This avoids the necessity of building individual delay units in each of the power units.

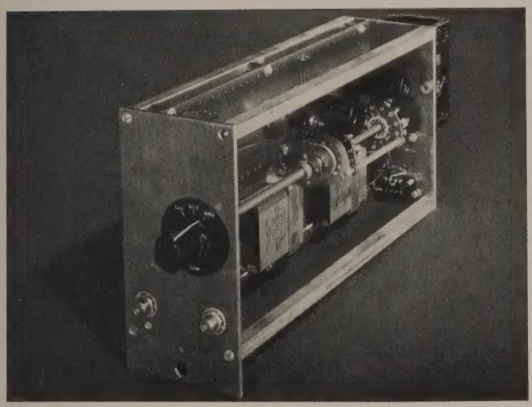

Fig. 1. High level pulse amplifier.

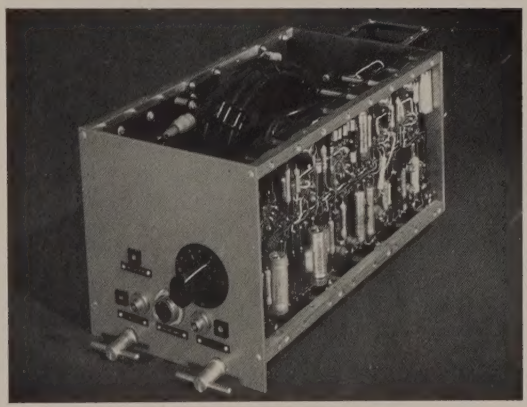

Fig. 2. Non-overloading pulse amplifier.

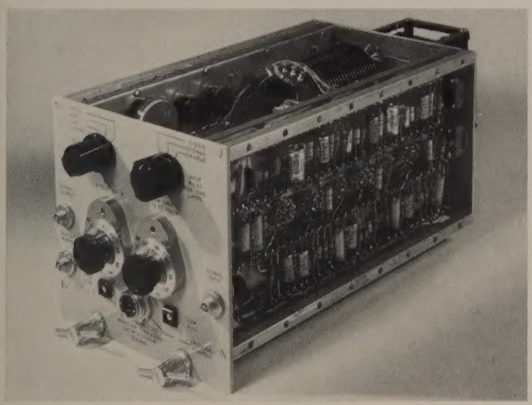

Fig. 3. Single channel pulse analyser.

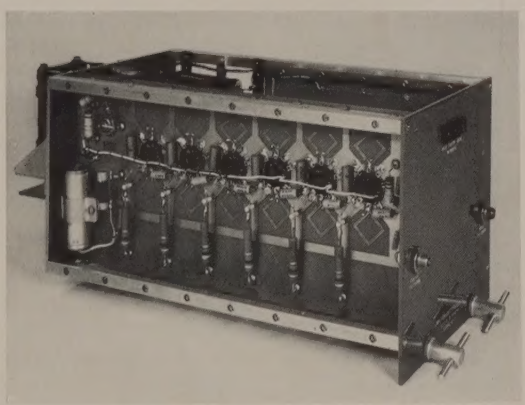

Fig. 4. Distributed amplifier.

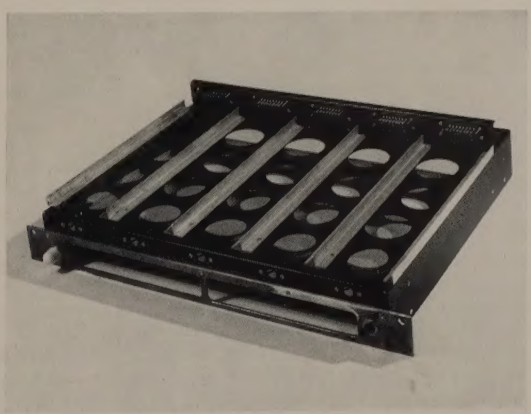

Fig. 5. Standard tray unit.

DISCUSSION

RISPOLI attached great importance to the standardisation of cable impedance. In Italy a distributed amplifier of 125 ohm impedance is at present being developed, and a cable impedance of 125 ohm had been chosen, as at CERN and at Berkeley. It appeared regrettable that Harwell and Saclay had however chosen 100 ohm.

ORMAN explained that one reason why 100 ohm impedance was preferred at Harwell was that satisfactory 100 ohm cable was readily available in the U.K., but 125 ohm cable was not. Although for the 125 ohm American cable RG 63-U there is listed an equivalent British cable UR 64, there are difficulties in manufacturing it in the U.K., related partly to the "copperweld" material for the inner conductor.

WELLS added that the difference between 100 ohm and 125 ohm could be thought of as a mere 20% of voltage, which did not seem great when considering photo-multiplier gains of the order of 10^7. A match can always be achieved at the expense of a small loss of voltage.

LANGEVIN said that at Paris (Faculté des Sciences, Laboratoire de Physique Nucléaire Orsay) they had drawn up a list of provisional specifications for instruments, which had much in common with the specifications prepared in other establishments. They did however differ in their choice of cable impedances. Originally, 75 ohm cable was adequate, but the advent of fast pulse techniques had led them to use 200 ohm cable with Hewlett Packard equipment. Then for certain fast coincidence circuits it was necessary to choose two cable impedances related in the ratio 1:2, and while at first 50 ohm and 100 ohm were selected, it was later desired to reduce this total range to two impedance values only. Thus they had finally chosen 75 ohm which was already well established, and 150 ohm, for which cable was readily available in France. Of the alternatives rejected, 100 ohm appeared unsuitable for fast circuits because of the increased power necessary, and increased price of equipment. Similarly 125 ohm shared the same disadvantages, and in addition was not easily available in France. Connectors for 150 ohm cable were available from French suppliers, provided quantities of 500 minimum were ordered. Existing 100 ohm connectors could easily be adapted by slight reduction of the inner conductor diameter.

VON DARDEL welcomed the fact that all parties seemed to want to use 75 ohm impedance. The choice between higher impedance values was difficult. Whilst a higher impedance permits larger pulse voltage amplitudes, this consideration is less important when improved photo-multiplier tubes with higher gain become available. In CERN it was decided that improvements in photo-multipliers justified a reduction from 200 ohm, but did not yet permit the change to 75 ohm. The intermediate value of 125 ohm was then chosen, since it facilitated the use of techniques developed at Berkeley. Eventual use of 75 ohm was favoured by the fact that there were available not only many cables of flexible types, but also cables of very high quality which could be used in the future for transmission of pulses with very short rise times of 10^{-9} s or 10^{-10} s.

FREEMAN added that for an air-spaced coaxial cable, of a given outer diameter, minimum attenuation is found for an impedance of 75 ohm approximately.

HUTCHINSON urged that attention should be paid to the problem of coaxial connectors for signal cables between instruments. If there are n different types of connector in use, and cables may be required to couple any one to any other one, then it is necessary to have a stock of n^2 different cable types. But if one standard type of connector is chosen, it is sufficient to have a reduced stock of only $2n$ different cable types, of which each cable carries the standard connector at one end, and one of the n different connectors at the other end.

MEUNIER described a modified form of the coaxial connector plug PL 259 (Type UHF) which was adapted to accept 100 ohm cable, but could still mate with normal Type UHF sockets on existing instruments. The reflection from this combination had not been measured.

WELLS claimed that reflections at the connector were not important, since the relevant frequency was not higher than about 500 MHz. It had been found that the mismatch of a 30 ohm connector on a 100 ohm cable could be tolerated.

MEUNIER said it was important to minimise the number of connectors, and to minimise the capacitance added at, for example, the anode of a phototube. But approximate matching appeared to be sufficient in most cases.

GATTI said that probably the only occasion when it was important to avoid reflections was when measurements were made of delay in a cable.

PIZER stated that in CERN efforts had been made to employ accurately matched connectors, particularly in conjunction with high-quality cables and with modern photo-multipliers. It was thought that matching would be important in many experimental systems.

VON DARDEL suggested that other establishments should send to CERN details of their practices in standardizing equipment, impedances, cables and connectors, with the reasons for their choice in each case. CERN could then compile further comparative tables for general distribution of this information. He urged that agreement should be sought on the policy of converging to 75 ohm when photo-multiplier developments permit.

KANDIAH supported these suggestions. The existing differences between impedances resulted from urgent construction programmes which could not be delayed until universal agreement was reached. But he thought that very few experiments would be handicapped by the discrepancies at present. For long-term purposes, however, general adherence to one standard value, probably 75 ohm, would become important, and any mismatches would become more significant.

RISPOLI added his general agreement, and said that although it might not be possible to obtain uniformity of practice in all laboratories, it was much more important that CERN should publish their own chosen standard practices. Then it would at least be possible for visiting physicists who come to CERN for experimental work to arrive prepared to meet these discrepancies.

PIZER drew attention to the existing list, mentioned in his paper, which gives some standards now in use in CERN.

GATTI proposed that it would also be useful to circulate lists of available commercial tapered lines, for impedance transformation at cable terminations.